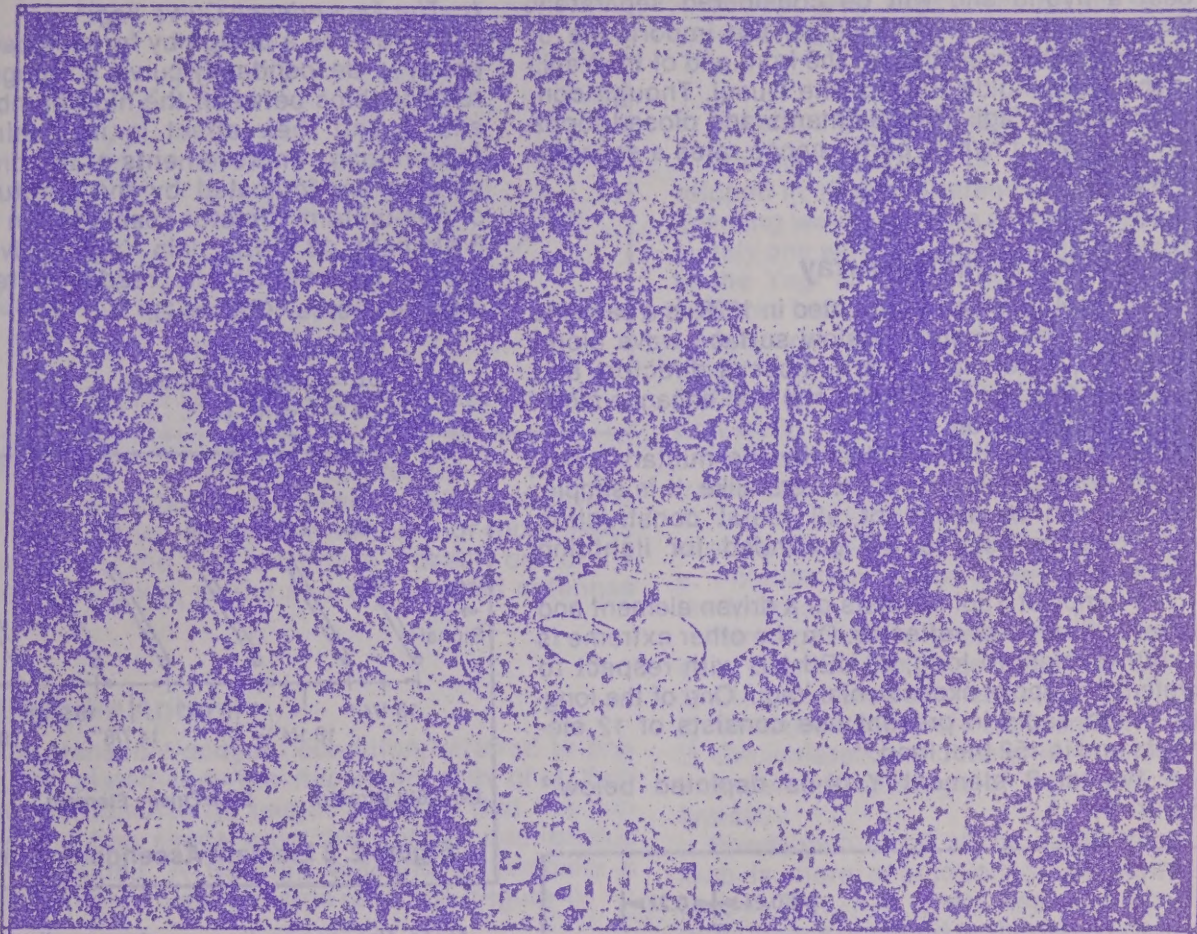


Antennas in the RF Range

... 10 kHz to 2000 MHz. A discussion of RF antennas used by all services.



Most antennas in the RF range fall into one of the following categories:

- A. Traveling wave
- B. Driven
- C. Vertical
- D. Leaky Waveguide
- E. Reflector

An antenna, as a transducer, converts electromagnetic energy into RF currents for receiving applications or launches an electro-magnetic wave from an RF source(s).

Though there is the appearance of five separate categories there is some degree of overlap just as there is overlap in the frequency domain and usage of all antennas.

Vertical antennas, especially when combined in arrays are not really a distinct group. They fall into the traveling wave or driven antenna category. How-

ever, especially in the lower end of the RF spectrum (10 kHz-30 MHz), vertical antennas are widely used and are discussed as a separate group.

Leaky wave antennas, such as the Directional Discontinuity Ring Radiator, or DDDR for short, though they launch and receive a vertically polarized wave and consequently could be grouped as "C", due to their unique construction are considered separately.

Traveling Wave Antennas

Type 1. This is a group of endfire antennas utilizing an array of parasitic elements or wave guiding structure with energy directed towards or from a primary driven element. The categories are:

- 1. Yagi-Uda Parasitic Array
- 2. Backfire
- 3. Quad or closed loop array

4. Quagi (Quad-Yagi)
5. Helical
6. Birdcage

Once again not all categories are distinct. A back-fire antenna is a Yagi and reflector combination. However since modifications are required to the Yagi when combined with the reflective screen it is at best a hybrid and will be considered separately. (This is not meant to diminish the importance of it firing off the back and not the front end of the Yagi.)

The same holds true for the Quagi. Though constructed of "Yagi type" elements and closed loops by virtue of its hybrid performance rates a separate category.

1. Yagi-Uda Parasitic Array

The Yagi-Uda antenna, invented in 1926, is a parasitic structure that supports a slow surface wave along its length. It consists of a number of directors and reflectors (the parasitic elements) that enhance radiation in one direction.

The Yagi has been utilized by the military, commercial interest and experimenters alike. It is simple, rugged, can be made of all metal construction and consequently be DC grounded for lightning protection.

The simplest Yagi consists of a driven element and either a director or reflector. On the other extreme is an 80 wavelength long, 26 dBd (dB with respect to a reference dipole) experimental Yagi.¹ One of the longest boom Yagis in present use consists of 12 elements and is 152 feet long.¹⁰

A typical 3 element Yagi is depicted below.² (Figure 1)

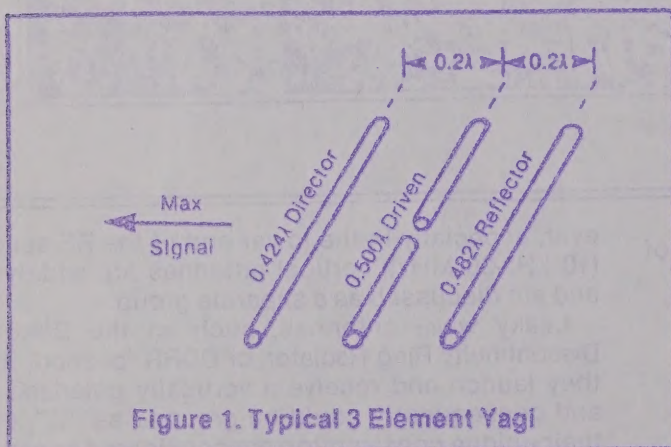


Figure 1. Typical 3 Element Yagi

Its elements can be linear, X-shaped or in the form of inverted Vees, each with its own merits.

a. Standard Design

There are several "recipes" for standardized construction of 3 or more element Yagis. One such recipe suggests that a single reflector be placed between 0.2 and 0.25 wavelengths behind the driven element and each added director of 2.5 percent diminishing length and 0.2 wavelength separation in front of the driven element.

The Hansen-Woodyard condition¹ specifies that high gains are achievable as long as the difference in

phase between the surface wave and the free space wave at the end of the antenna be equal to 180°. A general rule of thumb has always been "the longer it is the better it is (higher gain).

Recently several references have developed non-standard techniques for optimizing gain for a given boom length Yagi.

b. The Ascendency Curve Yagi³

Represents a design by formula called the ascendency — descendency curve. It recognizes the inter-relationships between the high number directors (5 and 6) and their effect in increasing the overall forward gain. Improvements of the order of 10 to 20 percent are indicated by the careful adjustments of non-uniform director *spacings*. An example of an 8 element, 13.1 dBd gain Yagi is given. For actual design information the reader is referred to the original source.³ (Figure 2).

D6 = 6th Director
DE = Driven Element
R = Reflector

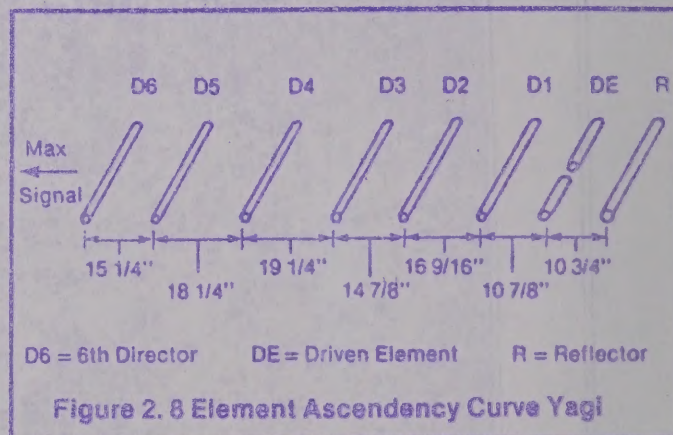


Figure 2. 8 Element Ascendency Curve Yagi

Note the ascending — descending spacings between the higher number directors. This Yagi is dimensioned in inches for operation at 222.5 MHz. (Which set of element lengths to use depends upon design bandwidth requirements and are clearly delineated in the reference.)

c. The Modulated Ladder Antenna

The modulated ladder antenna is basically a multi-element Yagi with non-uniform or "modulated" element lengths and spacings. The advantages indicated by this design are very high gain (26 dBd)¹ and minimum sidelobes over a substantial bandwidth. Properly designed modulated ladder arrays should exhibit an average beamwidth (H-plane beamwidth is larger, E-plane beamwidth is less) and gain as calculated from the formulas:

Beamwidth = $55\sqrt{l/\lambda}$ in degrees

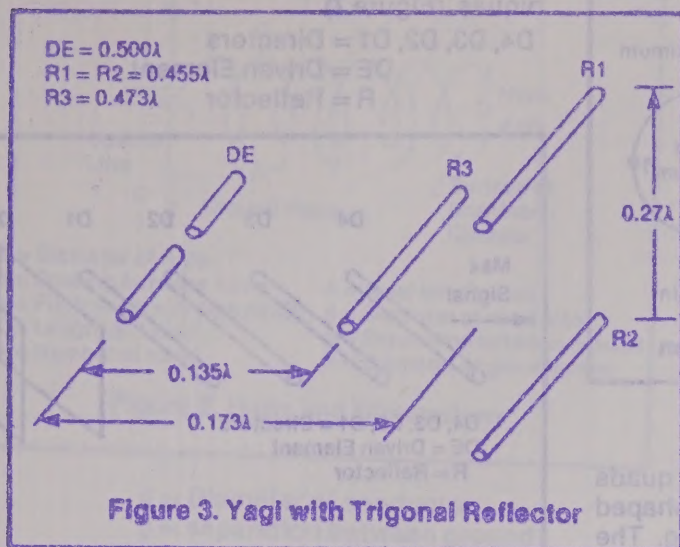
Gain = $10 \log_{10}(9l/\lambda)$ in dB above isotropic
 l = array length (same units as λ)

The concept relating higher achievable gains with non-uniform element lengths and spacings has been described in greater detail by David Cheng.^{4, 5}

d. The Trigonal Reflector

Most Yagi designs incorporate a single reflector. Experiments² utilizing plane reflecting surfaces, para-

bolas and corner reflectors were tried. The largest increase in gain (0.75 dB) occurred when a trigonal reflector was utilized as illustrated below in Figure 3.

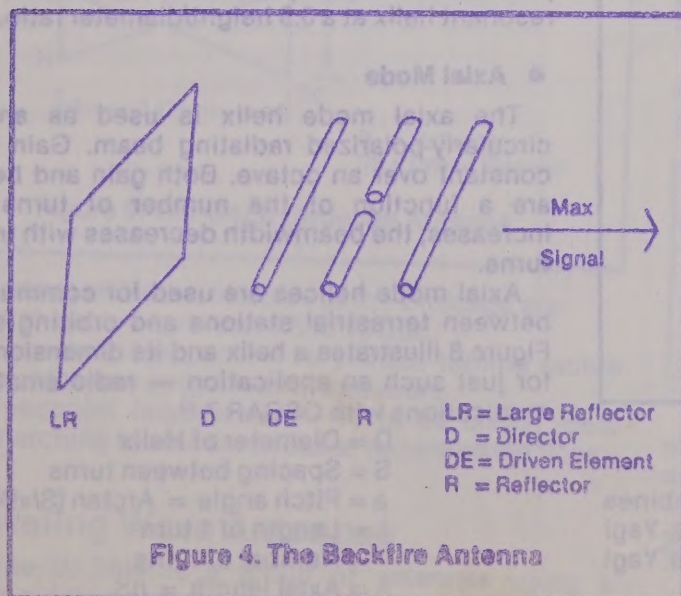


With well over 400 U.S. references to Yagi antenna development available since 1928⁶ and several definitive projects, some involving 3 or more years of research,² the Yagi antenna is probably one of the most interesting, useful and well studied antennas around.

2. Backfire Antenna

The Hansen-Woodyard condition, referred to previously, stated that the maximum gain of a Yagi is proportional to its length provided that the surface and free space wave 180° phase relationship is maintained. A Backfire antenna is basically a Yagi pointing towards a large reflector. (Figure 4)

LR = Large Reflector
D = Director
DE = Driven Element
R = Reflector



It takes advantage of the H-W condition by forcing the surface wave to traverse the physical length *twice*. This results in greater gain and frequency bandwidth than its equivalent length Yagi.⁷

Backfire antennas have been used at the high end of the RF range with gains in excess of 23 dB, at UHF with gains of 10.5 to 14 dB⁷ and at VHF with gain improvements of 4.5 dB over the isolated Yagi.⁸

3. Quad or Closed Loop Array

The cubical quad, or quad for short, was developed in 1942 by Clarence Moore.⁹ It is a parasitic traveling wave structure with loop elements approximately one wavelength in circumference. It is similar to the Yagi in several respects yet its differences are more widely known. These are:

- Approximately 2 dB greater gain than the equivalent length Yagi.
- "Quieter operation" — Less sensitive to precipitation static
- Feedpoint impedance closer in value to standard transmission line characteristic impedances
- Lower mounting height for the same takeoff or elevation angle as that of a Yagi.
- Polarization can be changed by simply changing feedpoints.

Handbook style design criteria for a 3 element quad follows from these basic formula:

- Circumference of driven element = 1005/f/MHz feet
 - Circumference of reflector element = 1030/f/MHz feet
 - Circumference of director element = 975/f/MHz feet
- Inter-element spacings of from 0.15 λ to 0.20 λ are typical.

Power gain in dB above an isotropic of a closed loop is more a function of its perimeter and not shape. This has given use to loop designs with other than a diamond or rhombus configuration.

• Circular Loops

Research on large circular loop arrays¹¹ have indicated that the loop perimeter (or enclosed area) and relative thickness of the conductors are gain determining factors. Antenna area and gain are related as follows:

$$G = 4\pi A/\lambda^2$$

where A (area) and λ² are entered with the same units of measure.

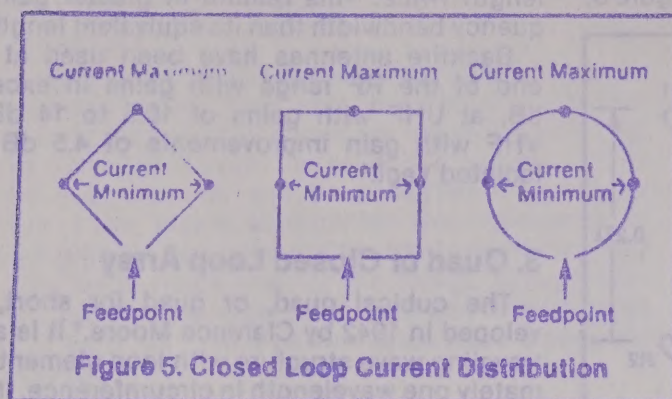
Figure 5 illustrates current distribution in diamond, square and circular 1 wavelength circumference loops:

• Delta Loops

Another popular form of loop array has a delta or triangular shaped element. It has the added advantage, especially at the lower end of the HF spectrum, in that it only requires a single high support, per element.

Delta loops exhibit substantially different take-off angles when inverted versus non-inverted, fed

at a corner or side and elevated versus erected close to the ground or ground system.^{12 13 14}

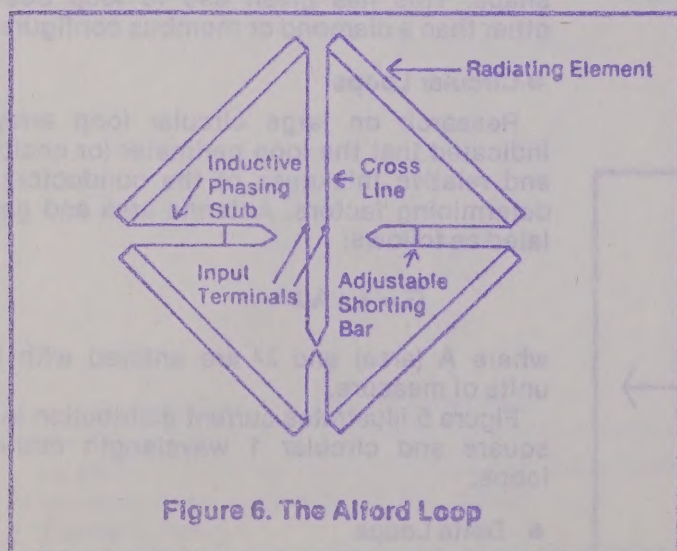


• Bi-square Array

As mentioned earlier conventional cubical quads have a 1 wavelength perimeter. A diamond-shaped quad would then have 4 sides each $\lambda/4$ long. The Bi-square loop is similar in shape and radiation direction (broadside) to the quad with the added advantage of 3 dB more gain. This additional gain is achieved by making the sides $\lambda/2$ in length. The total perimeter of a Bi-square loop is $4 \times \lambda/2$ or 2λ . By adding a parasitic reflector or director gains over 9 dBd have been claimed.¹⁵

• Alford Loop

One means of increasing antenna efficiency is to increase its radiation resistance. The Alford loop is a large square loop, consisting of four dipole radiators for sides. The radiation resistance is much higher than that of an ordinary loop. A typical Alford loop¹⁶ is indicated in Figure 6.

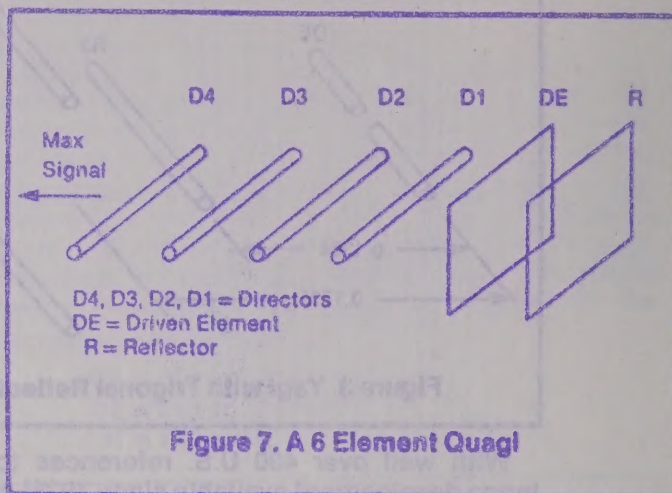


4. Quagis

The Quagi is a hybrid antenna that combines the best features of the cubical quad and the Yagi beam. Its directors are linear elements as in a Yagi while the driven and reflector are closed loops.

The immediately apparent advantages are greater gain over an equivalent number of element Yagi and simpler and more repeatable construction techniques. (Figure 7)

D4, D3, D2, D1 = Directors
DE = Driven Element
R = Reflector



5. Helicals

The front cover of this month's Issue depicts a right-hand circularly polarized helical antenna operating at 468 MHz. Unlike dipole antennas where the electro-magnetic wave travels at the velocity of light the velocity of the wave along the axis of the helix is lower. It depends on the frequency, diameter and number of turns per unit length.

Helical antennas are categorized by their shape (cylindrical, flat or conical) and their mode of operation: normal or axial.

• Normal Mode

Distinguishing characteristics of a normal mode helix are their small diameters and lengths with respect to wavelength. The radiated field is elliptically polarized. Circular polarization is obtained with a resonant helix at a 0.9 height/diameter ratio.¹⁷

• Axial Mode

The axial mode helix is used as an end-fire circularly-polarized radiating beam. Gain is nearly constant over an octave. Both gain and beamwidth are a function of the number of turns — gain increases, the beamwidth decreases with increasing turns.

Axial mode helices are used for communications between terrestrial stations and orbiting satellites. Figure 8 illustrates a helix and its dimensions utilized for just such an application — radio amateur communications with OSCAR 7.²⁶

D = Diameter of Helix
S = Spacing between turns
 $a = \text{Pitch angle} = \arctan(S/\pi D)$
L = Length of 1 turn
n = Number of turns
A = Axial length = nS

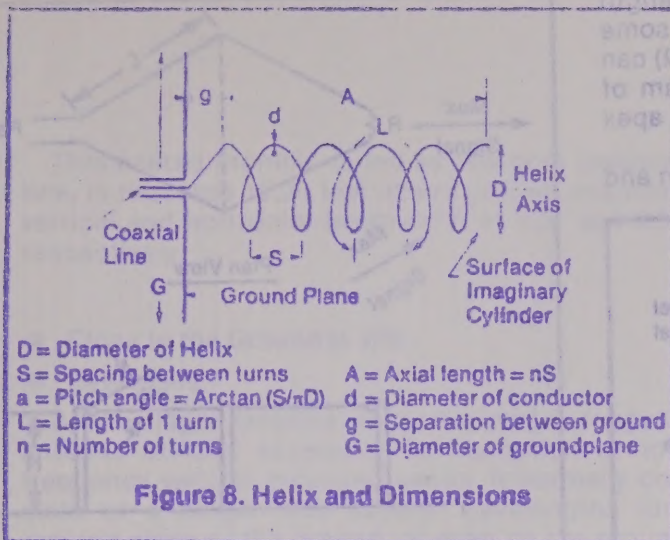


Figure 8. Helix and Dimensions

d = Diameter of conductor
 g = separation between ground
 plane and first turn
 G = Diameter of ground plane

6. Birdcage

The birdcage is a unique 2 element parasitic array that borrows favorable characteristics from the cubical quad and Vee beams. Invented and patented by Dick Bird, a British radio amateur, in 1958,¹⁹ it closely resembles a bird cage. (Figure 9)

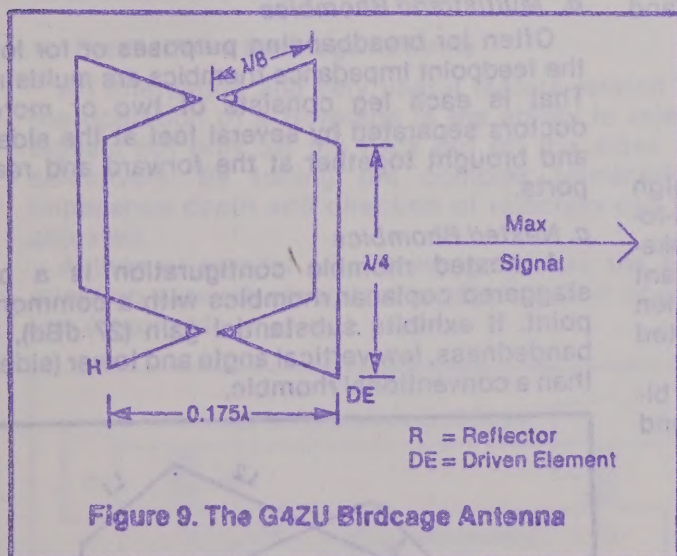


Figure 9. The G4ZU Birdcage Antenna

Several attractive features of the Birdcage are:

- High gain — between 8.5 to 10 dBd.
- Extremely compact — very small turning radius even at the low end of the HF spectrum.
- Feedpoint impedance — 40 to 50 ohms, ideal for matching to standard coaxial transmission lines.

Traveling Wave Antennas

Type 2. This is a group of antennas using a wave guiding structure, primarily wire, that is long

with respect to the operating wavelength. These antennas are additionally grouped by their height above ground.

• Elevated above ground ($>\lambda/8$)

- Long wire antennas
- Vee antennas, Vee beams
- Rhombics

Unlike the previous Type 1 (Traveling Wave Antenna) Type 2 antennas evolve. That is long wires "1" make up Vee antennas "2" and Vees combine to form Rhombics "3".

1. Long-wire Antennas

The old adage "the bigger the better" applies to this type of antenna. A long wire antenna when fed at one end either directly (Fuchs antenna) or by means of a transmission line has a radiation pattern that directly relates to its length and height above ground. Radiation from end-fed long wire antennas is broadside when the overall length is $5/8$ wave or less.

As the length of the long wire increases above $5/8$ wave the radiation pattern forms into a cone coaxial and away from the feed point as illustrated below. (Figure 10)

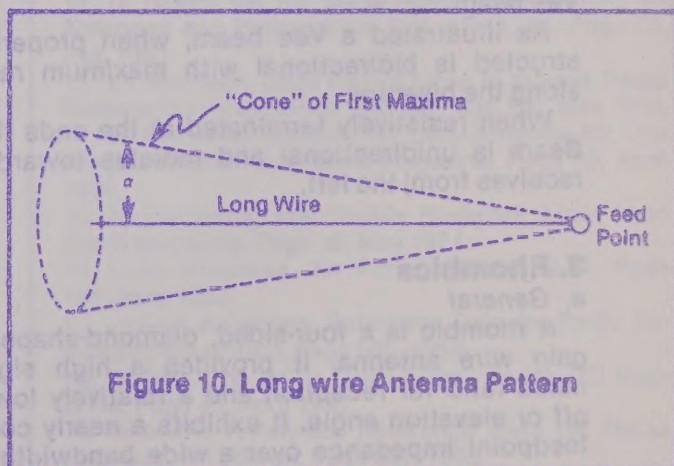


Figure 10. Long wire Antenna Pattern

A table of the locus (angle) of first maxima of the cone of radiation versus long wire length is given below: (Table 1)

Table 1			
α	λ	α	λ
53°	1	23°	5
37°	2	20°	6
29°	3	18°	7
26°	4	17°	8

The advantages of this type of antenna are its simplicity, high gain (for many wavelength long wires) and low cost to install and maintain.

2. Vee Antennas or Vee Beams

Vee antennas represent a natural progression from long wire antennas. They have the added ad-

vantages of greater gain, (3 dB versus same length long wires) sharper directivity and reduction of some sidelobes. Two long wires of specified length (λ) can be combined at one point to form a Vee beam of related angle (2α). This is to say that length and apex angle are related in optimized gain Vees.

Figure 11 illustrates spatial lobe combination and

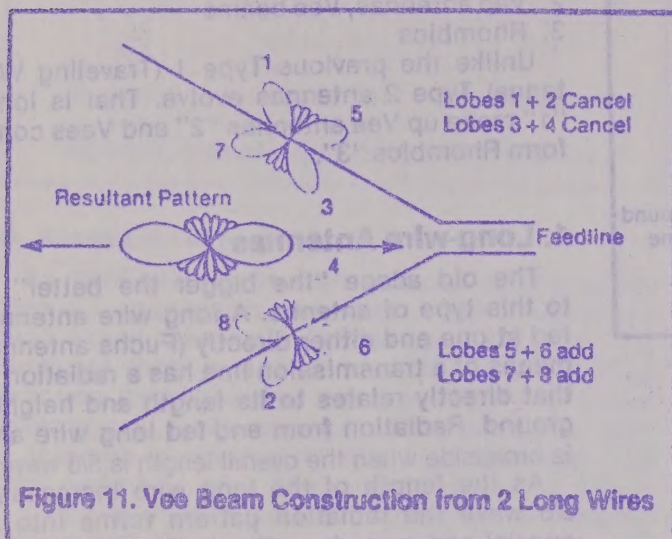


Figure 11. Vee Beam Construction from 2 Long Wires

cancellation in the long wire construction of a Vee beam.

As illustrated a Vee beam, when properly constructed is bidirectional with maximum radiation along the bisector.

When resistively terminated at the ends the Vee Beam is unidirectional and radiates towards (and receives from) the left.

3. Rhombics

a. General

A rhombic is a four-sided, diamond-shaped high gain wire antenna. It provides a high signal-to-noise ratio for reception and a relatively low take-off or elevation angle. It exhibits a nearly constant feedpoint impedance over a wide bandwidth when properly terminated at the far end. A terminated rhombic is illustrated below. (Figure 12)

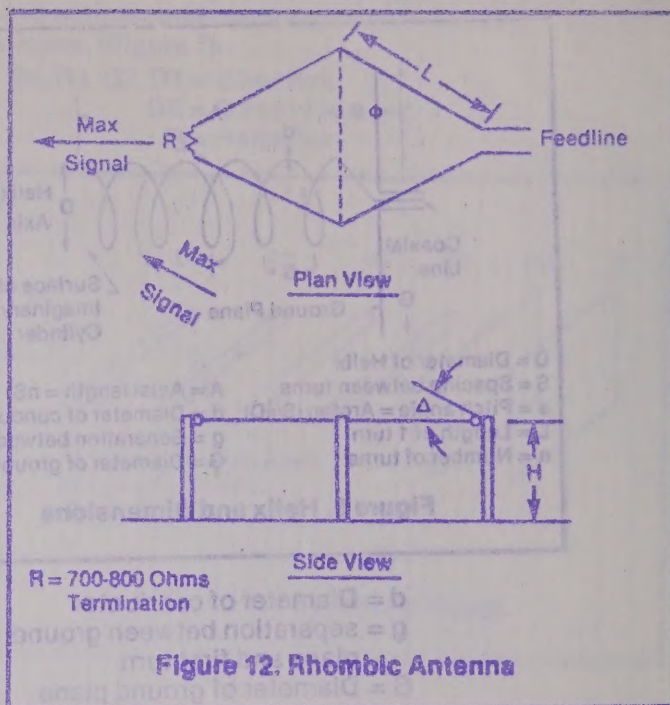
Maximum radiation is indicated along the bisector towards the left. Design terminology and criteria are as follows:

- ϕ = tilt angle
- Δ = vertical angle
- H = height above ground
- L = long side length

Using readily available charts¹⁰ and assigning values to two out of four of the parameters the remaining two are calculated.

Assume a 15° vertical angle is required. (Based on a specific propagation path and mode) and a maximum side length of 5λ is all the existing property will allow.

What tilt angle and height above ground does that calculate out to?



The referenced chart indicates that a tilt angle ϕ of 71° and a height of 0.95λ or 2.8λ will suffice. The lower of the two (heights) for practical considerations is chosen.

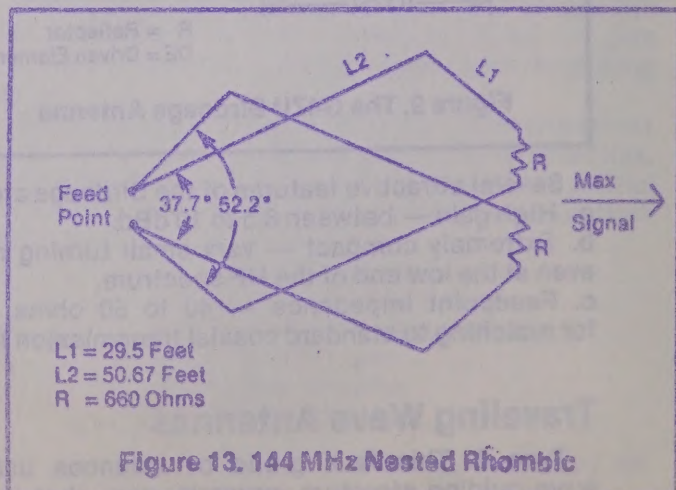
Terminations for rhombics are non-inductive resistors or in the case of very high power transmitters lossy balanced transmission line is utilized.

b. Multistrand Rhombics

Often for broadbanding purposes or for lowering the feedpoint impedance rhombics are multistranded. That is each leg consists of two or more conductors separated by several feet at the side poles and brought together at the forward and rear supports.

c. Nested Rhombics

A nested rhombic configuration is a pair of staggered coplanar rhombics with a common feedpoint. It exhibits substantial gain (27 dBd), broadbandedness, low vertical angle and lower (sidelobes) than a conventional rhombic.



L1 = 29.5 Feet
L2 = 50.67 Feet
R = 660 Ohms

This nested rhombic is fed by 300 ohm balanced line, is mounted 12.29 feet above ground and has a vertical and horizontal beamwidth of 5.5° and 8.5°, respectively.

- Close to the Ground ($\leq \lambda/8$)

- Beverages

A Beverage antenna is a horizontal longwire antenna utilized especially for reception of low-frequency vertical polarized waves. It normally consists of a single wire several wavelengths long erected close to the ground, or even on the ground, and terminated at the far end in its characteristic impedance.

An outstanding merit of this antenna is its ability to "pull signals out of the noise". The low end of the HF spectrum is typically atmosphere noise limited as opposed to thermal noise limiting in the higher RF ranges. Most HF antennas previously described present large signal and noise voltages at receiver terminals. The Beverage, on the other hand, generates a lower signal voltage while at the same time a considerably lower noise voltage. The overall effect is to produce a better or higher signal-to-noise ratio which is really the name of the game in communications intelligibility.

- Parallel Wire Beverage Antenna

An interesting and very useful feature related to the operation of a Beverage is the ability to reject signals not only off the back but off the sides — selectively. By tuning the complex termination impedance depth and direction of rejection can be adjusted.

A twin or parallel wire Beverage places the termination network close to the receiver (and operator). (Figure 14)

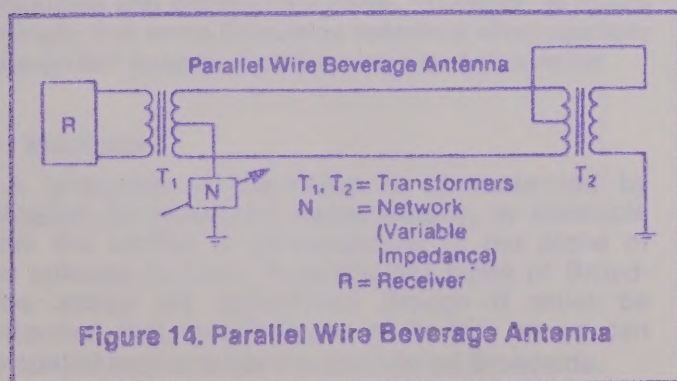


Figure 14. Parallel Wire Beverage Antenna

T_1, T_2 = Transformers
 N = Network (Variable Impedance)
 R = Receiver

Several comprehensive research efforts have explored many aspects of Beverage antenna design,

development, deployment and measurement.^{20, 21, 22, 23, 24} They are well worth reading about. Part 2 in the series continues with "Driven Antennas."

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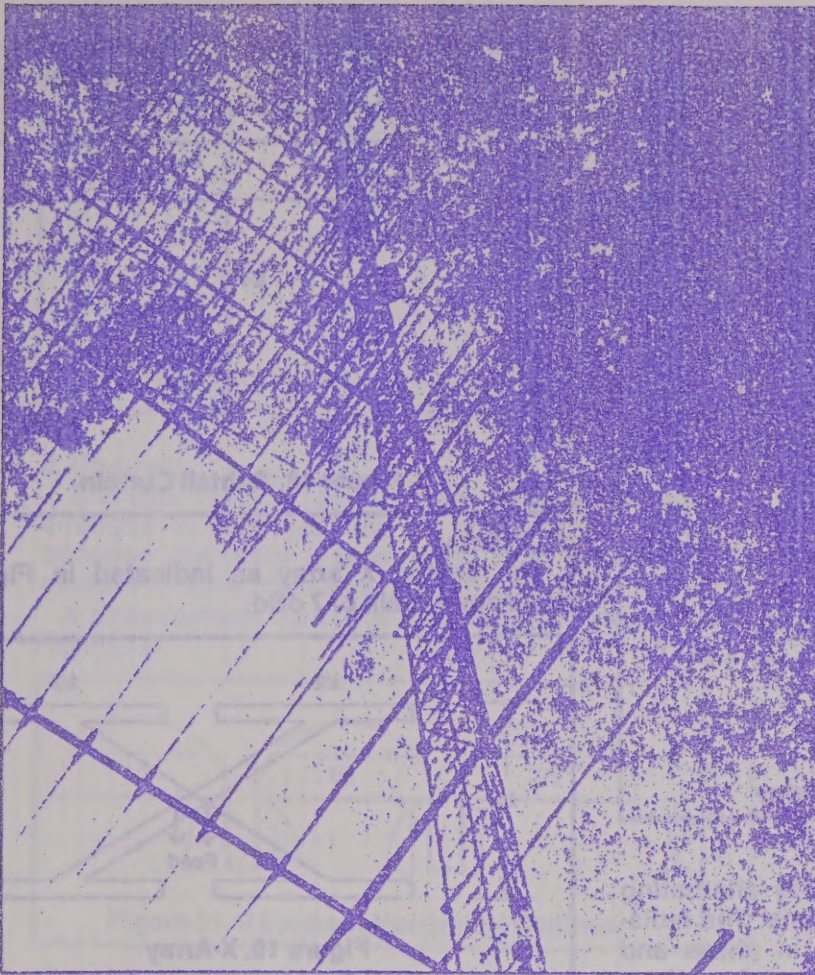


Photo of log periodic stacked array used in Artificial Radio Aurora experiments courtesy of Mr. Russell B. Stoner, N.T.I.A., Boulder, Colo.

Antennas In The RF Range

... 10 kHz to 2000 MHz. The discussion of RF antennas concludes with Driven, Vertical, Leaky-Wave and Reflector types considered.

Driven Antennas

A driven antenna can be defined as an array of two or more elements, separated in phase (electrical and/or spatial), hard connected via conductor or transmission line(s) and combined so as to develop a specific azimuthal or elevation pattern.

Driven antennas can be "loosely" subdivided into *Broadside* and *Endfire* categories. "Loosely" is meant to imply that some Broadside antennas when spatially rotated 90° become Endfire arrays and vice versa.

• Broadside

A broadside configuration is characterized by radiation (or reception) perpendicular, or steerable from the normal or perpendicular to the plane of the antenna or array. Four distinct types of Broadside arrays are considered though it must be reiterated that certain types of Endfire arrays can by spatial reorientation be considered Broadside.

The four types are:

1. Curtains
2. Dipoles, Collinears
3. Trombone T
4. Lazy H

Curtains

a. General

A curtain consists of an array of identical radiating elements configured in the vertical plane and in some cases backed by a reflecting screen.

Narrow beams and high broadside directivity gains are achieved by controlled amplitude and phasing techniques. Linear phase variations across the array enable the steering of transmitted or received wavefronts.

A uniformly spaced, uniform amplitude array has the advantage of producing the narrowest azimuthal beamwidth.

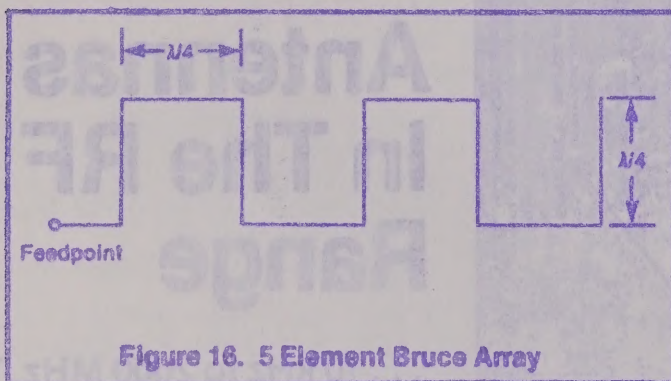
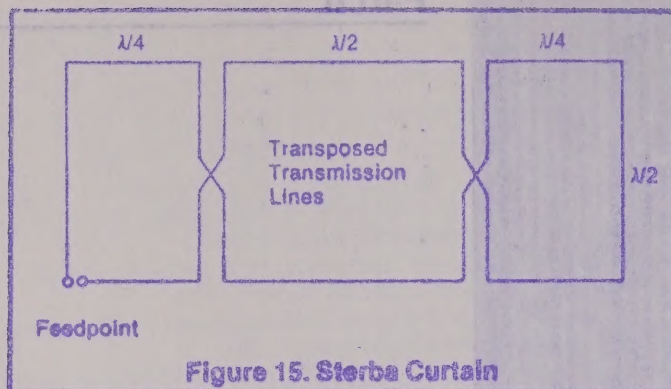
A uniformly spaced, tapered amplitude array such as a Binomial or Dolph-Tchebyshev distribution is utilized to minimize sidelobe level — a very useful technique in reducing susceptibility of a receiving antenna to interference.

b. Specific Designs

Each of the curtains illustrated below have substantial gain, various feedpoint locations and are inexpensive to build.

(1.) Sterba Curtain (Figure 15)

Probably the best known broadside array made up of a combination of parallel and collinear elements is the Sterba curtain.

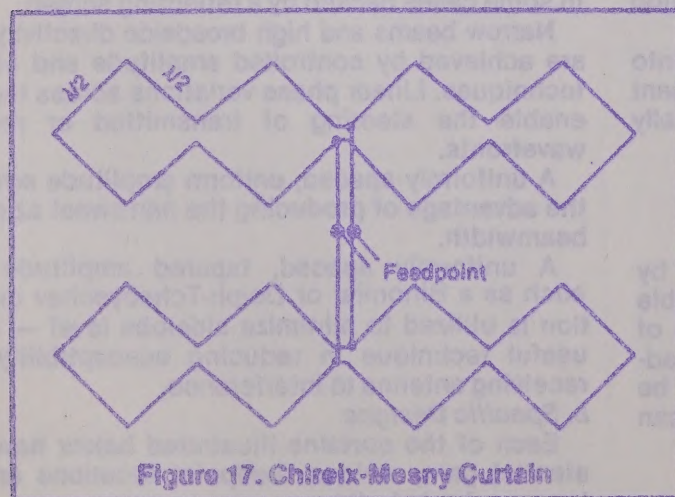


(2.) 5 Element Bruce Array (Figure 16)

The Bruce Array is configured with alternating $\lambda/4$ long vertical and horizontal elements. The radiation from the vertical sections are in phase and add while the radiation from the horizontal sections are out of phase and cancel. Typical gain of the 5 element array is 4 dB above a single quarter-wave vertical element.

(3.) The Chireix-Mesny Curtain

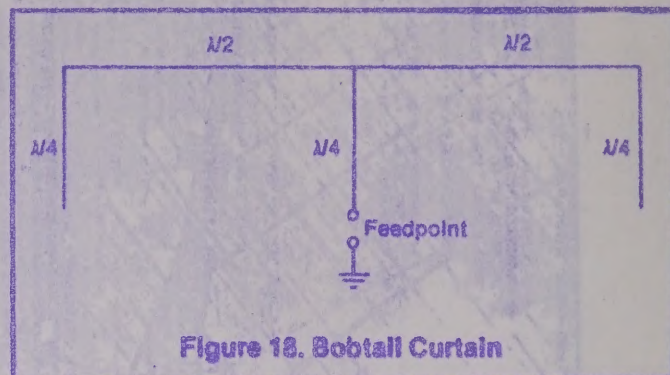
The C-M curtain consists of $\lambda/2$ dipoles configured in the form of diamonds. It is a center fed array with a broadside vertically-polarized radiation pattern. (Figure 17)



(4.) Bobtail Curtain

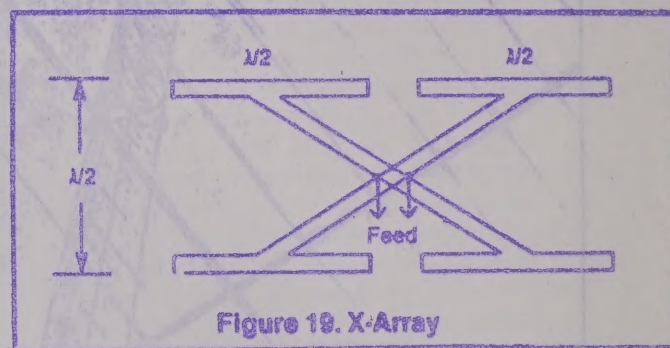
The Bobtail Curtain is a broadside array of three top current fed quarter wave verticals. Each vertical element is separated from the middle section by 180° . By virtue of only feeding the center vertical a

1:2:1 amplitude distribution occurs with its associated negligible sidelobe pattern. Theoretical gain is 4.9 dB over a single $\lambda/4$ vertical (Figure 18)



(5.) X-Array

The capital X array as indicated in Figure 19 has a power gain of 7 dBd.



Dipoles and Collinears

a. Unstubbed

There are two distinct categories of unstubbed dipoles: center and off-center fed.

(1.) Center-fed Dipoles

Flat-tops and Inverted-vees. Both configurations and important properties are tabulated below by length:

Length	Name	Gain
$\lambda/2$	half-wave	0 dBd*
λ	double Zepp	1.8 dBd
1.28λ	extended double Zepp	3.0 dBd

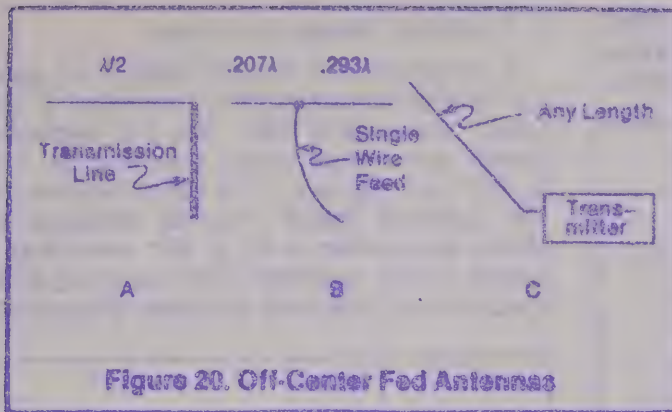
*(This is the reference dipole)

The extended double Zepp is the longest center-fed dipole with a radiation pattern remaining at right angles to the line of the antenna. (Max. energy).

(2.) Off-center Fed

Antennas and their characteristics are tabulated below:

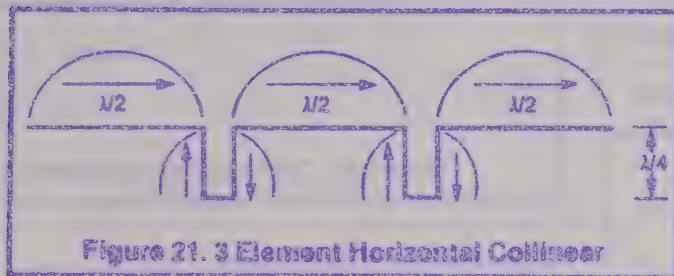
Antenna	Figure	Characteristics
Zepp	20-A	Voltage fed with balanced transmission line
Windom	20-B	Resonates on even order harmonics
Fuchs	20-C	Long wire brought directly to transmitter



b. Stubbed

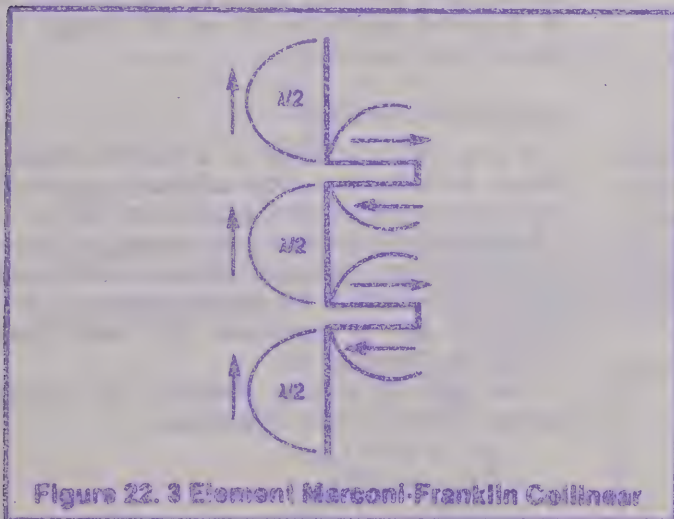
Stubbed "dipoles" are in-phase radiating elements arranged in a single line or collinear. The stub between elements insures that each successive element is fed at the same phase relationship.

A three element horizontal collinear is illustrated in Figure 21.



It is a simple antenna and can be fed at almost any point. Additional gain is achieved simply by adding another stub and half-wave element.

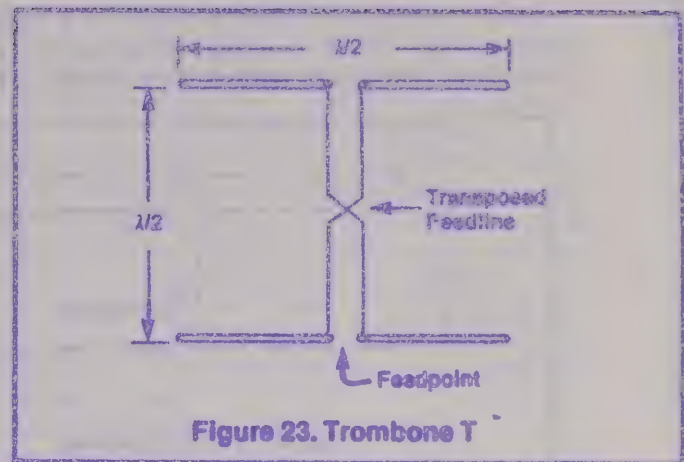
The vertical counterpart of this collinear is called a Marconi-Franklin array. (Figure 22)



In both antennas current direction is indicated by the arrows.

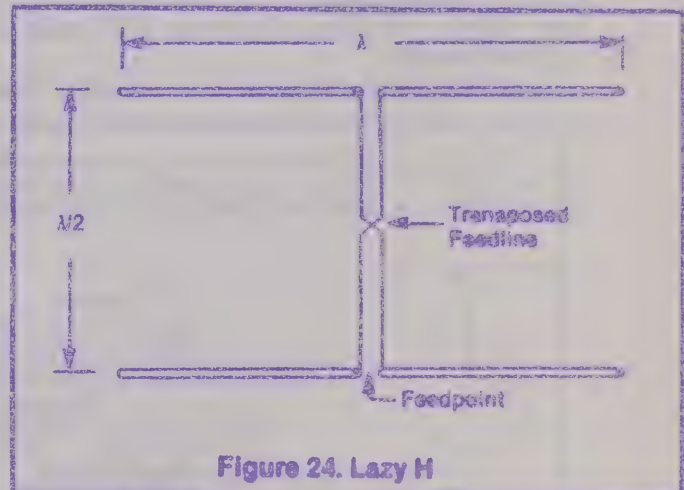
3. Trombone T

The Trombone-T is a simple 4 dBd gain broad-side antenna of low-cost. It is illustrated in Figure 23.



4. Lazy H

The Lazy-H is a twice as large version of the Trombone T. Gain is approximately 6 dBd (5.8 dBd) and it is voltage instead of current fed. See Figure 24.



● Endfire

An endfire configuration is characterized by radiation in the same plane formed by the antenna or array. It can be further subdivided into the broad categories of Frequency Dependent or Frequency Independent.

● Frequency Dependent

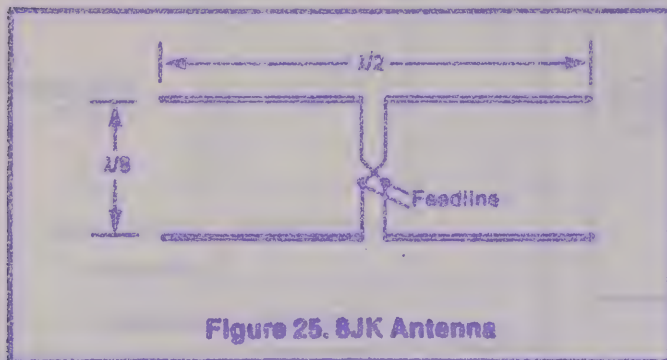
Antennas that operate over a narrow band i.e. maintain directional patterns and feedpoint impedance response appropriate for its intended use are considered in the classification -- Frequency Dependent. They are:

1. 8JK
2. Double Triplex Beam
3. ZL Special
4. Cardioid Beam Antenna

1. 8JK

The 8JK was named after its inventor John D. Kraus W8JK (Radio amateur call sign). It is an

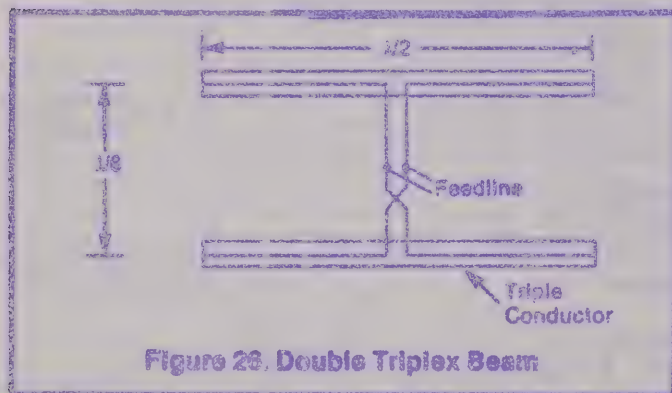
endfire array with approximately 4.3 dBS bidirectional gain. It is a simple antenna to construct with reliable performance. It is illustrated in Figure 25.



2. Double Triplex Beam

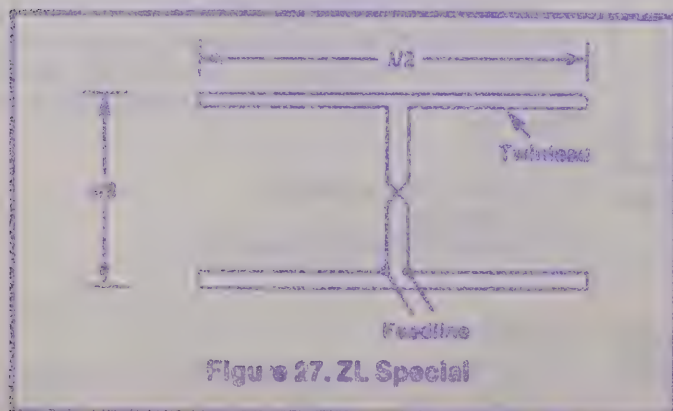
The Double Triplex Beam is basically an 8JK with the single linear conductors replaced by triple leads. It has the added advantages over the single 8JK of:

- Broadband frequency response
- High radiating efficiency (300 ohm feedpoint impedance)
- Good wet weather performance. (See Figure 26.)



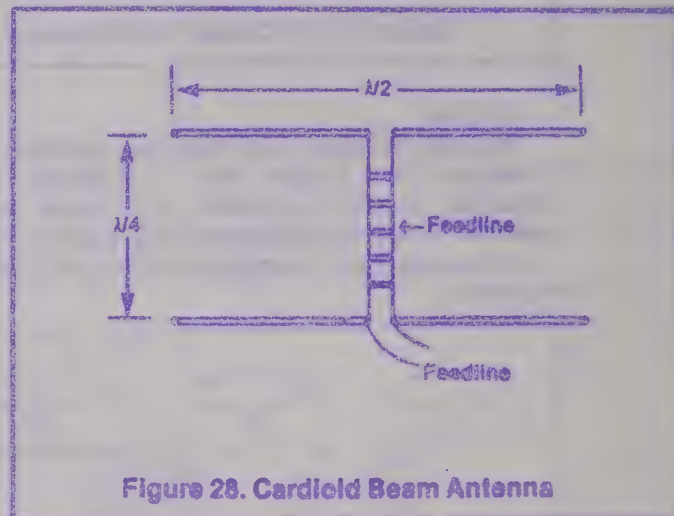
3. ZL Special

The ZL Special is a two element driven end-fire array. The elements are electrically driven 135° out of phase and are spatially phase separated by an additional 45°. It is unidirectional, has a 20 dB front to back ratio and is lightweight and easy to construct. (Figure 27)



4. Cardioid Beam Antenna

A unidirectional endfire pattern is generated if the total-phase separation (electrical & spatial) between elements is 180°. (i.e. the respectively generated wavefronts must combine in 1 direction (0, 360°) and cancel in the opposite direction (180°). The Cardioid Beam Antenna develops this 180° phase separation with a 90° electrical and 90° spatial phase condition. The resulting pattern is of a broad-nosed cardioid response. (Figure 28)



● Frequency Independent

Conversely, antennas that maintain a relatively constant terminal (feedpoint) impedance and vertical angle (elevation angle) or display other frequency invariant directional properties, over several octaves, are classified as frequency independent. The logarithmically periodic array or log periodic for short is a prime example of an RF range frequency independent antenna.

Log Periodic Arrays²⁰

A Log Periodic array is a multi-element end-fire driven antenna in which adjacent elements length and spacing are equal to a constant ratio τ .

A second parameter σ is related to the ratio of the separation between adjacent elements and 4 times the length of the shorter element. An illustration and set of formulas will help clarify. See Figure 29. (On page 15).

The gain of a Log Periodic is a function of τ and σ . Other design criteria are:

$$2l_1 = 0.47 \lambda_{\max}$$

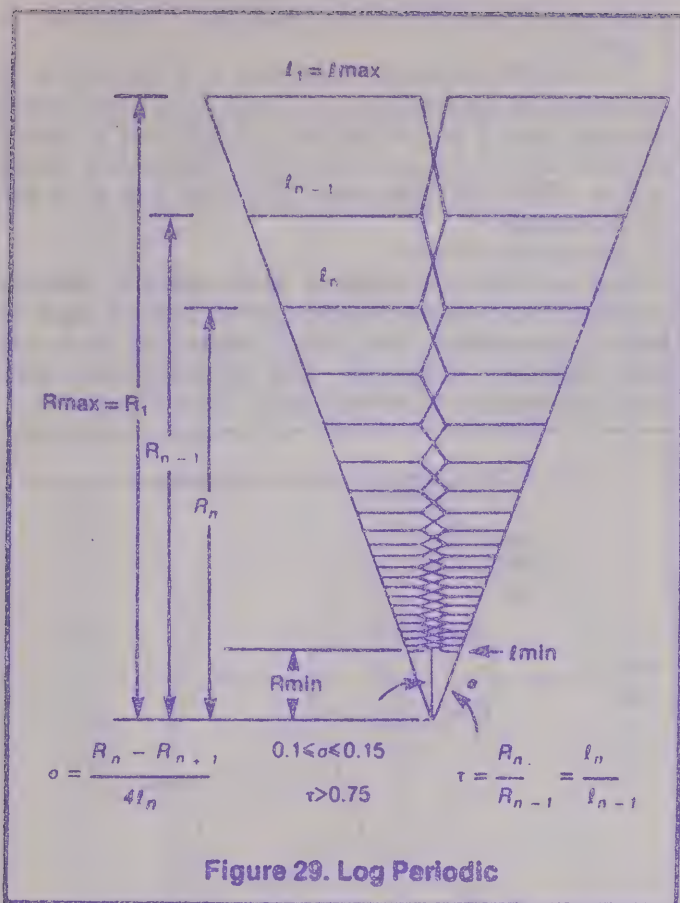
$$2l_{\min} \leq 0.38 \lambda_{\min}$$

$$R_1 \cdot R_{\min} \geq 0.50 \lambda_{\max}$$

With these criteria and the required high and low frequencies 3 nomographs are used to complete the design.

Verticals

The simplest of vertical antennas is sometimes represented as the evolution of an open-ended



coaxial transmission line. The outer conductor is folded back exposing the inner conductor and the combination of inner and outer conductors form a complete radiating system or antenna with specific feedpoint impedance and directional pattern.

The vertical antenna is probably the most widely recognized radiator by the public due to its prolific utilization by commercial AM broadcast stations, naval whips and hobbyists in the form of amateur and CB car and roof-mounted ground planes.

Discounting re-oriented horizontal antennas and special categories such as the stubbed collinear or Franklin array, discussed earlier, verticals can be classified as follows:

1. Single driven conductor, length $\leq 5/8 \lambda$
2. Single driven conductor, length $> 5/8 \lambda$
3. Broadbanded verticals
4. Special application verticals

1. Single Driven Conductor, Length $\leq 5/8 \lambda$

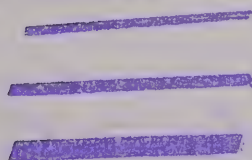
Vertical antennas are commonly utilized at the following heights: $1/4 \lambda$, $1/2 \lambda$ and $5/8 \lambda$. All three heights lay down a maximum signal strength (theoretically) at the horizon or zero elevation angle. This requires that the ground mounted vertical work against an infinite extent, infinite conductivity ground system.

The field strength, under these conditions, along the horizontal as a function of antenna height for

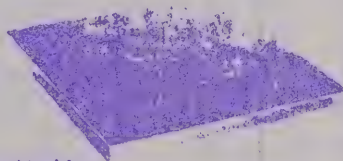
MICROWAVE ABSORBER

Rigid, Flexible & Castable Materials for your -

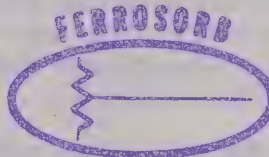
- Coaxial & Waveguide load elements
- Attenuator elements
- Back Cavity Spurious Absorbers
- Harmonic Absorber Filters
- Rotary Joint Gaskets



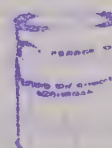
Machinable
Round & Square Bars
Waveguide ID Bars



Machinable
Plate Stock
1/8" - 1" TK & 3x3 to 12x12



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a vertical grounded radiator with one kilowatt radiated power is tabulated below:

Antenna Height	Field Strength
$1/4\lambda$	196 mv/meter
$1/2\lambda$	236 mv/meter
$5/8\lambda$	276 mv/meter

Such is rarely the case. Usually, in commercial service, (AM broadcast) 120 radials are laid out every 3 degrees. This does help stabilize the impedance of the vertical system though in itself does not lower the maximum radiation angle to the horizon.

Simple techniques are however available for establishing an extremely low (as low as 0°) elevation or takeoff angle. Mounting a vertical above ground with a radial or counterpoise system drooping downward at about 45° helps to significantly lower the elevation angle (of maximum radiation). One precaution worthwhile mentioning is that the counterpoise should be at least a $1/4\lambda$ above ground.

Numerous studies have developed working relationships between vertical antenna efficiency, length and number of radials.^{26, 27, 28, 29}

2. Single Driven Conductor, Length $> 5/8\lambda$

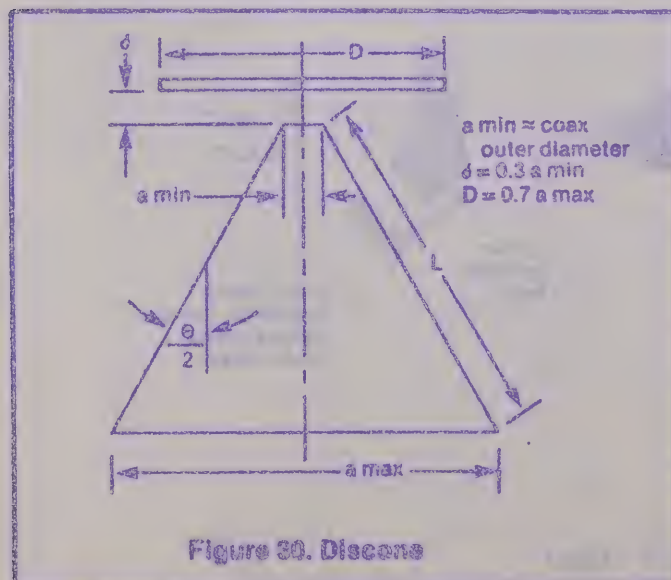
Verticals greater than $5/8\lambda$ in continuous length are difficult to construct for the lower frequencies and as a result of vertical angle lobe splitting and its concomitant high angle radiation are not as widely utilized as the previous category.

3. Broadbanded Verticals

There are several vertical antennas that operate over large bandwidths with VSWRs not exceeding 2.0:1. Two of them, the Discone and the Conical Monopole are discussed.

a. Discone

The discone is a wideband vertical radiator that can be directly matched to a 50 ohm system. The

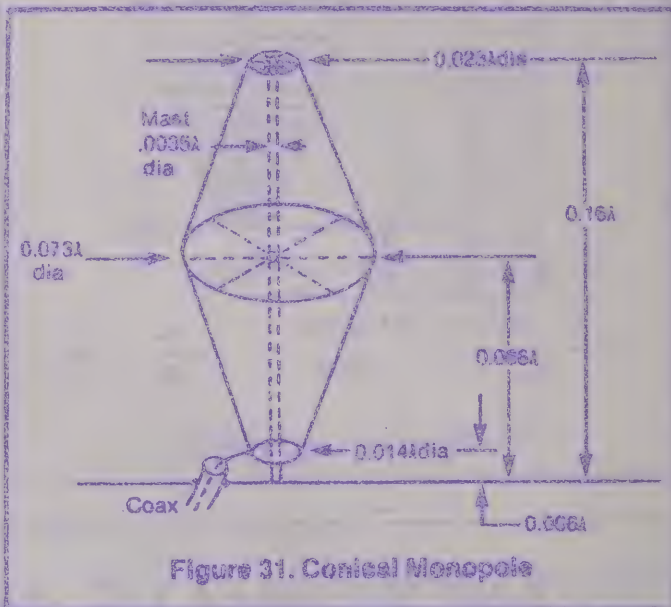


basic configuration of a Discone is indicated in Figure 30.

The VSWR frequency variation at the low end of the band is determined by L and θ . An optimized discone with $\theta = 60^\circ$ and $a/L = 1/22$ has a VSWR of less than 1.5:1 over a 7 to 1 frequency range and a VSWR of less than 2.0:1 over a 9 to 1 frequency range.³⁰

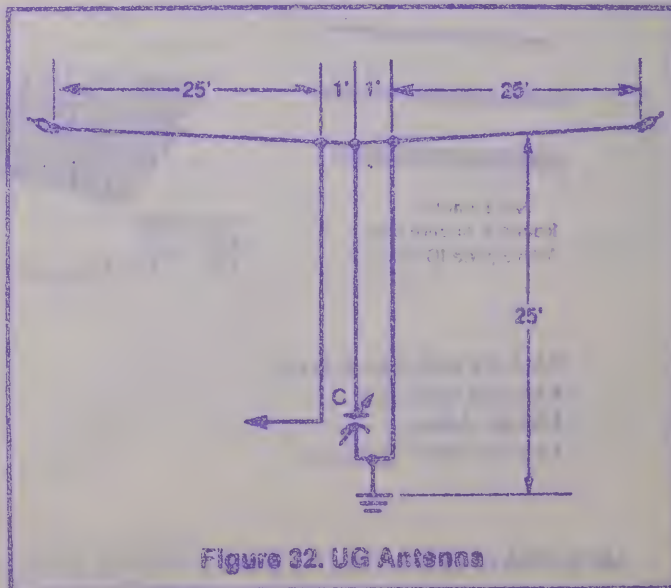
b. Conical Monopole

The conical monopole is short vertical radiator resembling two terminated cones placed base to base. It operates over a 4:1 bandwidth, and has high radiation efficiency. The configuration with dimensioning is indicated below.³¹ (See Figure 31)



4. Special Application Verticals

LF and VLF applications are probably the hardest areas to accommodate in terms of antenna design due to several conflicting constraints. Foremost is the operating wavelength — in a word — long. Reasonable feedpoint impedances, bandwidth and



efficiencies must be maintained. Two, out of many configurations, that are in themselves tradeoffs, are discussed.

a. Type UG

The type UG antenna has the shape of a capital T as illustrated below with dimensioning given for 2 MHz operation. Capacitor C is varied until a favorable feedpoint impedance is developed.⁹²

(Figure 32)

b. NORD

The NORD is a short vertical antenna, used at LF and MF. It has a sufficiently high bandwidth and radiation efficiency for its intended use. It is basically three over-coupled tuned circuits. A center tower is used as a common element. See Figure 33.

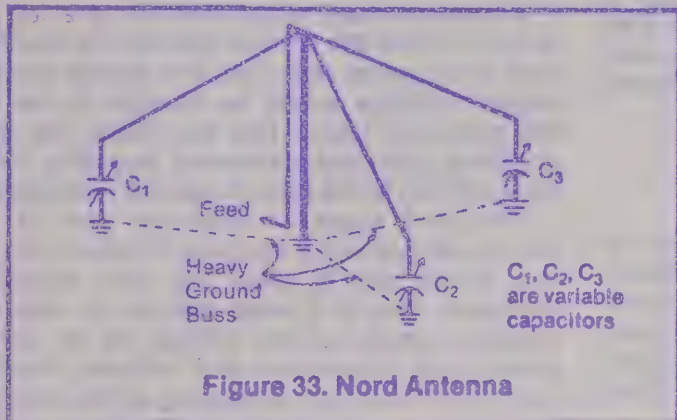


Figure 33. Nord Antenna

Leaky Wave Antennas

Leaky wave antennas can be identified by their low silhouette (close to ground) or controllable aperture distribution i.e. beamwidth and sidelobe level. Several types identified here are:

1. DDRR
2. Fast Wave
3. Slow Wave

1. DDRR

a. $\lambda/4$ DDRR

The Directional Discontinuity Ring Radiator developed by J.M. Boyer is an omni-directional low profile vertical polarization radiator. It is a single circular open-ended element mounted horizontally a short distance above a ground plane. See Figure 34.

According to theory the DDRR antenna ring forms a transmission line with its ground plane. The slot, between the ring and ground plane being a discontinuity, radiates continuously in a direction transverse to the ring axis.

The advantages of this antenna include small horizontal dimensions, extremely low height and simple construction. Since it is DC grounded induced charge static is not a problem as it is with a monopole.

b. $\lambda/2$ DDRR

A larger version of the original DDRR is available. Primary advantage is an increase in gain. It is as depicted in Figure 35.

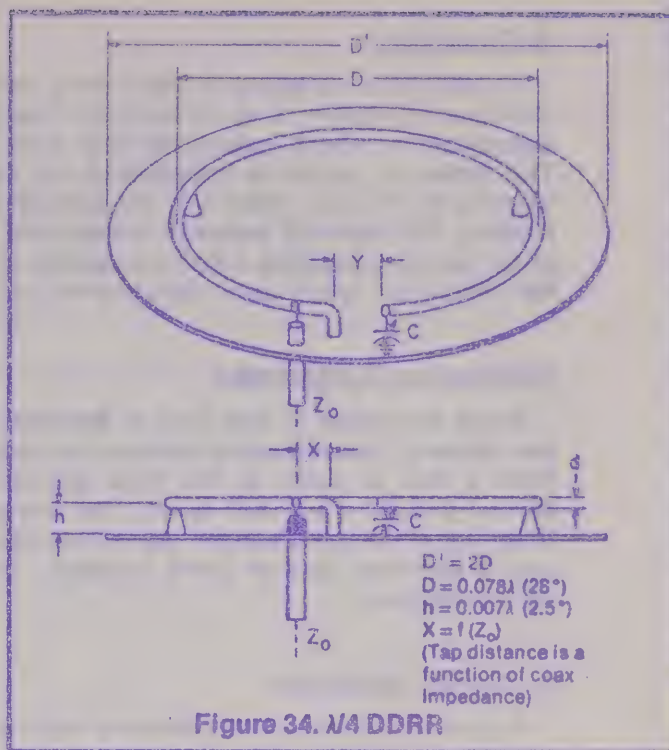


Figure 34. $\lambda/4$ DDRR

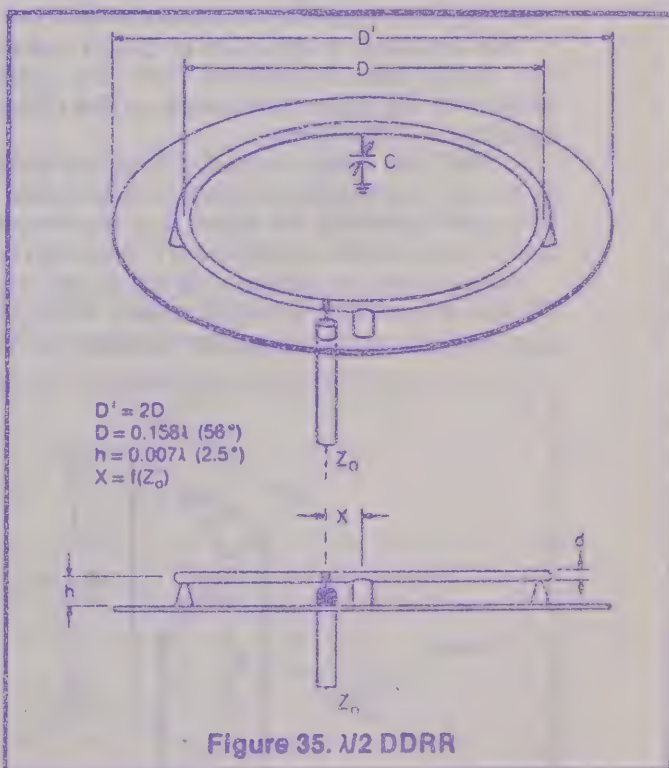


Figure 35. $\lambda/2$ DDRR

2. Fast Wave

Another class of leaky wave antenna is the Fast Wave Structure. Similar to the DDRR it is a low silhouette structure. It is cut along its length to permit the radiation of energy⁹⁴. The angle of radiation is closely related to the phase velocity of the guiding structure before the cut is made. The gaps must be smaller than $\lambda/3$ of the highest frequency of operation.

3. Slow Wave

An additional category of leaky wave antenna is a slow-wave structure. It is periodically-loaded (series capacitors or shunt inductors) long wire antenna. The radiation angle is determined by the phase velocity of the slow wave and the periodicity of the loading. The spacing between successive loadings must be approximately a half-wavelength or greater, for with closer spacings no radiation occurs.

Reflector Antennas

Beam formation in this type of antenna relies on the shape of the reflecting surface to focus energy from a line or point at the feed position into a narrow parallel wavefront beam. The converse also holds i.e. intercepted parallel wavefronts are focused into a common line or point location. Two types considered are:

1. Corner Reflector

A corner reflector is a half-wave radiator with a reflector consisting of two plane metal surfaces meeting at an angle immediately behind the radiator.

The radiation is the sum of direct radiation from the dipole and reflections from the planes. Consequently the pattern depends on the distance from the dipole to the corner.

At VHF the corner reflector is often solid (metal). At lower frequencies screen wire or metal rod constructed reflectors are used. The maximum gap between the wires or rods should not exceed 0.06λ .

Power gains of at least 10 dB to as much as 14 dB are achievable using a corner reflector. A diagram with design information is shown in Figure 36.

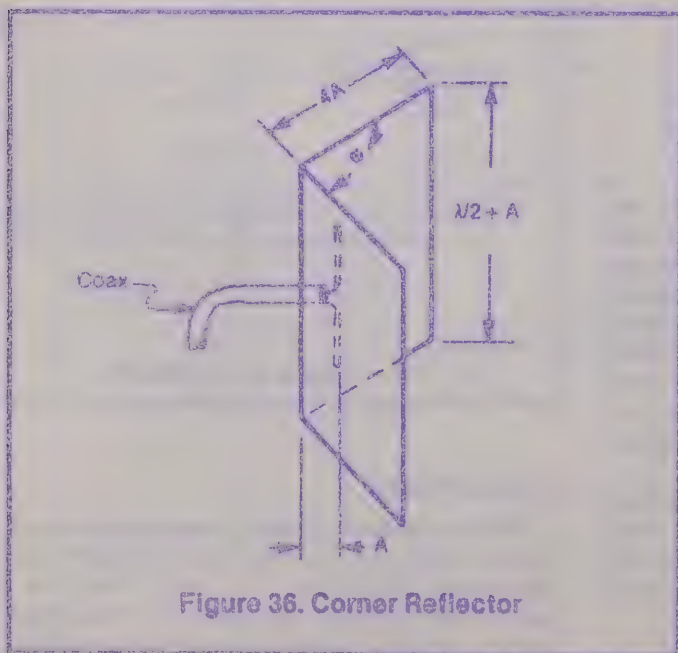


Figure 36. Corner Reflector

Principal advantages of the corner reflector are its simple construction and the low side and back lobes of the directivity pattern.

2. Parabolic Reflectors

Parabolic reflectors are widely used in UHF and microwave radar and point-to-point communications systems to achieve antenna gains in excess of 20 dB.

Recently the CATV industry has been successfully utilizing a Half-bollic to receive very weak over the horizon VHF signals.³⁴ (A Half-bollic, as its name implies is physically half a parabolic reflector. When combined with its ground image its performance is similar to a parabolic reflector.)

In the low frequency application an offset feed is employed to reduce undesired interaction between reflector and feed and to minimize aperture blocking.

A suitable feed might consist of a log periodic array in which the longest element forms the dipole feed of a corner reflector. The phase center of the corner reflector would be located at the focus of the parabola. Using this technique the maximum possible gain and narrowest possible beamwidth are achieved at the lowest operating frequency.

Though 5 types of antennas with 51 different configurations were discussed this is not to imply that these are, or were the only antennas ever designed. This is a broad sample of most common antenna configurations utilized by all services — commercial, government and hobbyist. Modifications such as shortening, employing active elements and material experimentation affects the properties and probably the classifications of the antennas already discussed.

Future articles will delve into specialized antenna configurations utilizing some of the above-mentioned techniques, and others not yet considered. □

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Radiation

Introduction

Figure 1 shows an elemental antenna structure connected to an ac source. We wish to determine the conditions under which such a structure will effectively radiate electromagnetic energy into the surrounding region, and we wish to develop analytical relations for describing this radiation. We have

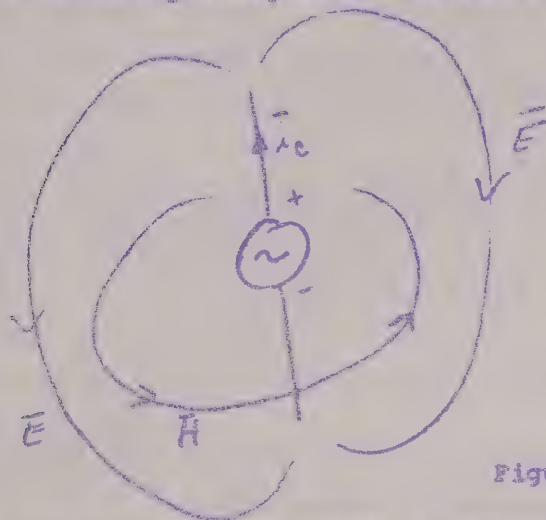


Figure 1

seen earlier that for static fields, we can view the current as the source of the magnetic field and the charge as the source of the electric field. We then can determine the fields from knowledge of the sources most effectively by the use of intermediary potentials. In the dynamic case the electric and magnetic fields become coupled, and hence a knowledge of one is tantamount to a knowledge of the other since they are always related to each other by Maxwell's equations. We may also conclude that a knowledge of one source distribution is sufficient to determine the fields, and such is indeed the case.

This conclusion is also consistent with another point of view, namely that because of the continuity relation, charge and current cannot be independently specified and hence the two source distributions that were independent for statics are now necessarily coupled. Since in many practical situations the current is known or can be approximated, we pick the current as the most efficacious choice of the source and develop relations by which the fields can be calculated from arbitrary current distributions.

Potential Theory

Earlier, in the discussion of electrostatics, we have seen that

$$\vec{E} = -\nabla\phi \quad (1)$$

where ϕ is the electrostatic potential function. Since, by a vector identity also used earlier,

$$\nabla \times \nabla \phi \equiv 0 \quad (2)$$

we conclude that Eq. 1 cannot hold in the dynamic case because

$$\nabla \times \nabla \phi = -\nabla \times \bar{E} \neq 0 \quad (3)$$

We postulate that a time dependent term is missing in Eq. 1 and, in addition, that the potential ϕ should be replaced by the time dependent potential $\phi(t)$, which, when time variation ceases, reduces to ϕ . Thus we rewrite Eq. 1 for the dynamic case as

$$\bar{E} = -\nabla \phi(t) + \bar{Y}(t) \quad (4)$$

where for no time variation,

$$Y(t) \rightarrow 0, \quad \phi(t) \rightarrow \phi \quad (5)$$

Now, performing the operation of Identity 2, the result is

$$\nabla \times \bar{E} = -\frac{\partial \bar{B}}{\partial t} = \nabla \times (-\nabla \phi(t)) + \nabla \times \bar{Y}(t)$$

or

$$\nabla \times Y(t) = -\frac{\partial \bar{B}}{\partial t} \quad (6)$$

In order to determine $Y(t)$ in Eq. (6), we relate $\bar{B}$ to a potential.

Guided by our experience with magnetostatics where, since

$$\nabla \cdot \bar{B} = 0 \quad (7)$$

and for any vector,

$$\nabla \cdot \nabla \times \bar{A} \equiv 0 \quad (8)$$

we wrote

$$\bar{B} = \nabla \times \bar{A} \quad (9)$$

where in (9), $\bar{A}$ is the magnetostatic vector potential, we recognize that for dynamic fields, eqs. 7 and 8 still hold, so for the dynamic magnetic field we write

$$\bar{B} = \nabla \times \bar{A}(t) \quad (10)$$

where $\bar{A}(t)$ is the dynamic vector potential, and for no time variation

$$\bar{A}(t) \rightarrow \bar{A} \quad (11)$$

Using Eq. 10 in Eq. 6, there results

$$\nabla \times Y(t) = -\frac{\partial}{\partial t} (\nabla \times \bar{A}(t))$$

or

$$Y(t) = -\frac{\partial \bar{A}(t)}{\partial t} \quad (12)$$

This result applied to Eq. 4 finally gives the electric field (and hence the magnetic field using Maxwell's equations) from a knowledge of the potential. The

$$\bar{E} = -\nabla \phi(t) - \frac{\partial \bar{A}(t)}{\partial t} \quad (13)$$

Vector Potential, $\vec{A}(t)$

While Eq. 13 is useful to us in the derivations that follow, it is not the best equation to use for finding the fields simply because it involves both potentials, and from our earlier considerations, we know that the potentials are coupled and knowledge of one should be sufficient. Indeed if we use Eq. 10 then a knowledge of $\vec{A}(t)$ alone is sufficient to determine $\vec{B}$, and through Maxwell's equations, $\vec{E}$. Hence, this latter procedure is the one to be followed. How then do we determine $\vec{A}(t)$ from an arbitrary current distribution?

In order to find a differential equation relating $\vec{A}(t)$ to $\vec{I}(t)$, we start with the Maxwell equation

$$\nabla \times \vec{H} = \vec{I}_C + \epsilon \frac{\partial \vec{E}}{\partial t} \quad (11)$$

and cast it in terms of the potentials by substituting from Eqs. 10 and 13. This results in

$$\nabla \times \frac{1}{\mu} \nabla \times \vec{A}(t) = \vec{I}_C - \epsilon \frac{\partial}{\partial t} \left[\nabla \phi(t) + \frac{\partial \vec{A}(t)}{\partial t} \right] \quad (12)$$

By considering μ to be independent of position, i.e., not a function of xy and using the vector identity,

$$\nabla \times \nabla \times \vec{A} = \nabla(\nabla \cdot \vec{A}) - \nabla^2 \vec{A} \quad (13)$$

Eq. 12 can be rewritten as

$$\nabla \left(\nabla \cdot \vec{A}(t) + \mu \epsilon \frac{\partial \phi(t)}{\partial t} \right) - \nabla^2 \vec{A}(t) = \mu \vec{I}_C - \mu \epsilon \frac{\partial^2 \vec{A}(t)}{\partial t^2} \quad (14)$$

Recalling from earlier notes that a vector field is completely defined when both its curl and divergence are specified, we turn our attention to the vector field $\vec{A}(t)$. To this point we have defined $\nabla \times \vec{A}(t)$ through Eq. 10, namely

$$\nabla \times \vec{A}(t) = \vec{B} \quad (15)$$

Thus only the rotational source has been specified. To completely specify the vector field $\vec{A}(t)$ we must define its divergence and are free to do so in any way we choose since $\vec{A}(t)$ is an arbitrary vector. Hence, let us take

$$\nabla \cdot \vec{A}(t) = -\mu \epsilon \frac{\partial \phi(t)}{\partial t} \quad (16)$$

The specification of the irrotational source of $\vec{A}(t)$ through 16 is called the Lorenz gauge condition or simply the Lorenz condition. Use of the Lorenz condition (16) in Eq. 12 results in the desired equation for $\vec{A}(t)$

$$\nabla^2 \vec{A}(t) - \mu \epsilon \frac{\partial^2 \vec{A}(t)}{\partial t^2} = -\mu \vec{I}_C \quad (17)$$

We recognize Eq. 19 as a wave equation for $\bar{A}(t)$. That is, the solutions of the homogeneous equation in (19) are propagating waves with phase velocity $1/\sqrt{\mu\epsilon}$. We view the term $\mu\bar{I}_C$ as the forcing function for this equation, and indeed we have already interpreted physically the current as the source of the potential and hence the fields. One notes that for the static case, Eq. 19 reduces to

$$\nabla^2 \bar{A} = -\mu \bar{I}_C \quad (20)$$

the familiar equation for the magnetostatic vector potential. The solution of Eq. 20 is known from previous experience to be

$$\bar{A}(\mathbf{r}) = \iiint \frac{\mu \bar{I}_C(\mathbf{r}') dV}{4\pi |\mathbf{r} - \mathbf{r}'|} \quad (21)$$

where $\bar{r}$ and $\bar{r}'$ are defined in Fig. 2. Thus to determine the total vector potential at point $\bar{r}$, one must add the contributions of each source current element

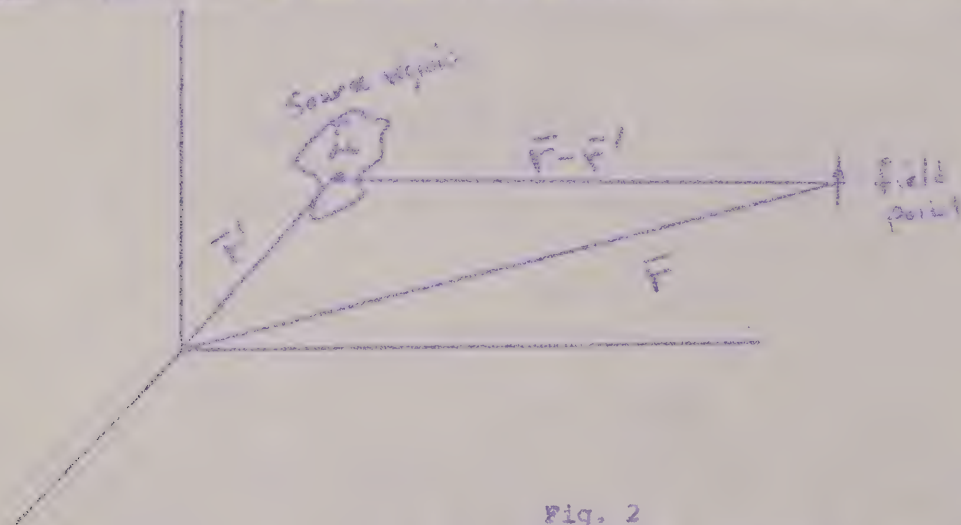


Fig. 2

$\bar{I}_C$ by integration over the source volume. Guided by 21 and knowledge of the solution to the homogeneous equation in 19, we write the solution to Eq. 19 as

$$\bar{A}(\mathbf{r}, t) = \iiint \frac{\mu \bar{I}_C \left(\mathbf{r}' - \frac{|\mathbf{r} - \mathbf{r}'|}{v_d} \right) dV}{4\pi |\mathbf{r} - \mathbf{r}'|} \quad (22)$$

where

$$v_d = \frac{1}{\sqrt{\mu\epsilon}}.$$

The potential given by 22 is called the retarded vector potential. This can be readily understood if one recalls the knowledge gained from earlier study of point potentials in transmission lines. One recognizes the term in the integrand as a propagating wave function, and the interpretation of 22 is that the potential at $\mathbf{r}$ is affected by changes in the current at $\mathbf{r}'$ only after a time

delay of $|\vec{r}-\vec{r}'|/v_\phi$ seconds. Of course for harmonic time variation, Eq. 22 becomes

$$\vec{A}(\omega, \vec{r}) = \iiint \frac{\mu \vec{I}_c(\vec{r}') e^{-j\omega |\vec{r}-\vec{r}'|}}{4\pi |\vec{r}-\vec{r}'|} dV' . \quad (23)$$

The procedure to be followed in finding the radiation fields from a harmonically varying current distribution is then as follows. Given the current, determine $\vec{A}(\vec{r})$ from 23 by integrating over the source. Use $\vec{A}(\vec{r})$ in Eq. 10 to determine the magnetic field and finally the Maxwell Eq. 14 to obtain the electric field.

Scalar Potential, $\phi(t)$

Although we already have sufficient information for the solution of any free space radiation problem, we will develop the relations for the scalar potential, $\phi(t)$, in the interest of completeness and hopefully for a better overview of potential theory. We now seek a differential equation linking $\phi(t)$ with the charge distribution, ρ . In statics this relation is Poisson's equation, so our result must reduce to Poisson's equation when time variation ceases. We start with the Maxwell equation from Gauss' law

$$\nabla \cdot \vec{E} = \rho/\epsilon \quad (24)$$

and substitute for $\vec{E}$ from Eq. 13 which gives

$$-\nabla \cdot \left(\nabla \phi(t) + \frac{\partial \vec{A}(t)}{\partial t} \right) = \frac{\rho}{\epsilon}$$

or

$$-\nabla^2 \phi(t) - \frac{\partial (\nabla \cdot \vec{A}(t))}{\partial t} = \frac{\rho}{\epsilon} . \quad (25)$$

Now using the Lorentz condition 18, Eq. 25 becomes

$$\nabla^2 \phi(t) - \mu\epsilon \frac{\partial^2 \phi(t)}{\partial t^2} = -\frac{\rho}{\epsilon} \quad (26)$$

as the desired result. Again the propagating wave nature of $\phi(t)$ is evident with the charge as the forcing function. We also note that for the static case Eq. 26 does indeed reduce to Poisson's equation. From our recent experience we expect the solution of Eq. 26 to be

$$\phi(t) = \iiint \frac{\rho \left(t - \frac{|\vec{r}-\vec{r}'|}{v_\phi} \right)}{4\pi\epsilon |\vec{r}-\vec{r}'|} dV' \quad (27)$$

where

$$v_\phi = \frac{1}{\sqrt{\mu\epsilon}}$$

Summary

The complete equations of potential theory are summarized in Table 1.

Table 1

Equations	General Time Variables	Harmonic Time Variation
$\vec{B} = \nabla \times \vec{A}$	$\vec{B} = \nabla \times \vec{A}(t)$	$\vec{B} = \nabla \times \vec{A}(t)$
$\nabla \cdot \vec{A} = 0$	$\nabla \cdot \vec{A}(t) = -\mu\epsilon \frac{\partial \phi(t)}{\partial t}$	$\nabla \cdot \vec{A}(t) = -j\omega\mu\epsilon \phi(t)$
$\vec{E} = -\nabla\phi$	$\vec{E} = -\left(\nabla\phi(t) + \frac{\partial \vec{A}(t)}{\partial t}\right)$	$\vec{E} = -\nabla\phi(t) - j\omega\vec{A}(t)$
$\nabla^2\phi = -\frac{\rho}{\epsilon}$	$\nabla^2\phi(t) = -\mu\epsilon \frac{\partial^2\phi(t)}{\partial t^2} = -\frac{\rho}{\epsilon}$	$\nabla^2\phi(t) + \omega^2\mu\epsilon\phi(t) = -\frac{\rho}{\epsilon}$
$\nabla^2\vec{A} = -\mu\vec{I}_C$	$\nabla^2\vec{A}(t) = \mu\epsilon \frac{\partial^2\vec{A}(t)}{\partial t^2} = -\mu\vec{I}_C$	$\nabla^2\vec{A}(t) + \omega^2\mu\epsilon\vec{A}(t) = -\mu\vec{I}_C$
$\phi(x) = \iiint \frac{\rho(x')}{4\pi\epsilon \vec{x}-\vec{x}' } dv$	$\phi(x,t) = \iiint \frac{\rho(x') \epsilon \left(t - \frac{ \vec{x}-\vec{x}' }{v_\phi}\right)}{4\pi\epsilon \vec{x}-\vec{x}' } dv$	$\phi(x,t) = \iiint \frac{\rho(x') \epsilon^{-j\omega \frac{ \vec{x}-\vec{x}' }{v_\phi}}}{4\pi\epsilon \vec{x}-\vec{x}' } dv$
$\vec{A}(x) = \iiint \frac{\mu\vec{I}_C}{4\pi \vec{x}-\vec{x}' } dv$	$\vec{A}(x,t) = \iiint \frac{\mu\vec{I}_C \epsilon \left(t - \frac{ \vec{x}-\vec{x}' }{v_\phi}\right)}{4\pi \vec{x}-\vec{x}' } dv$	$\vec{A}(x,t) = \iiint \frac{\mu\vec{I}_C \epsilon^{-j\omega \frac{ \vec{x}-\vec{x}' }{v_\phi}}}{4\pi \vec{x}-\vec{x}' } dv$

Radiation from a Current Element

Let us now apply the concept of the vector potential to determine the radiation fields of a current element. A current element is a current carrying conductor that is very short compared to a wavelength. One could imagine such an element to be, for example, a short circuit at the end of a balanced transmission line. This case is shown in Fig. 3 along with an appropriate coordinate system. From consideration of Fig. 2 we see that for this example $\alpha' = 0$ and

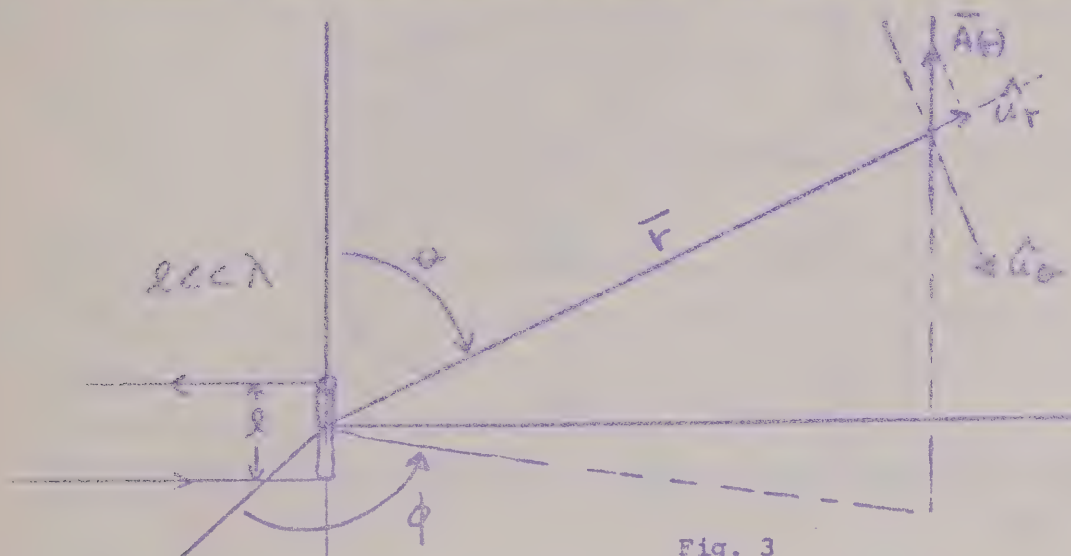


Fig. 3

hence $\bar{r} - \bar{r}' = \bar{r}$. We consider the driving transmission line to produce harmonically varying currents in the element and hence for our example $\bar{A}(\mathbf{r})$ is given by Eq. 23. For this case Eq. 23 becomes

$$\bar{A}(\mathbf{r}) = \iiint \frac{\mu \bar{\mathbf{i}}_c(\mathbf{r}') e^{-j\beta|\bar{\mathbf{r}}|}}{4\pi |\bar{\mathbf{r}}|} dV$$

Integrating over the source current volume yields

$$\bar{A}(\mathbf{r}) = A_z(\mathbf{r}) \hat{\mathbf{u}}_z = \frac{\mu I l e^{-j\beta r}}{4\pi r} \hat{\mathbf{u}}_z$$

where I is the total current in amperes and $|\bar{\mathbf{r}}|$ is simply designated as r .

The factor $I l$ is called the current moment. One should bear in mind that 29 applies to the special case of uniform current and $l \ll \lambda$. In addition, one should note that for rectangular coordinates, $\bar{A}(\mathbf{r})$ has the direction of $\hat{\mathbf{u}}_z$.

The electromagnetic field created by the current element can now be found by the procedure stated earlier. First we use Eq. 10 to find the magnetic field, i.e.,

$$\bar{\mathbf{H}} = \nabla \times \bar{\mathbf{A}} = \left(\frac{\mu I l e^{-j\beta r}}{4\pi r} \right) \hat{\mathbf{u}}_\phi$$

We would like to express the fields in spherical coordinates for convenience, and hence in order to implement the calculation indicated by Eq. 30, $\bar{A}(r)$ must also be expressed in spherical coordinates. From reference to Fig. 3 one sees

$$A_\phi = 0, \quad A_r = A_z \cos\theta, \quad A_\theta = -A_z \sin\theta \quad (31)$$

and the expression for curl $\bar{A}$ in spherical coordinates is given by

$$\nabla \times \bar{A} = \begin{vmatrix} \frac{\hat{u}_r}{r^2 \sin\theta} & \frac{\hat{u}_\theta}{r \sin\theta} & \frac{\hat{u}_\phi}{r} \\ \frac{\partial}{\partial r} & \frac{\partial}{\partial \theta} & \frac{\partial}{\partial \phi} \\ A_r & r A_\theta & r \sin\theta A_\phi \end{vmatrix} \quad (32)$$

Substitution of 31 into 32 and the use of Eq. 30 yields

$$\bar{E} = \frac{\mu I l}{4\pi} \begin{vmatrix} \frac{\hat{u}_r}{r^2 \sin\theta} & \frac{\hat{u}_\theta}{r \sin\theta} & \frac{\hat{u}_\phi}{r} \\ \frac{\partial}{\partial r} & \frac{\partial}{\partial \theta} & \frac{\partial}{\partial \phi} \\ \frac{e^{-j\beta r}}{r} \cos\theta & -e^{-j\beta r} \sin\theta & 0 \end{vmatrix} \quad (33)$$

Carrying out the operations in 33 results in only one component for the magnetic field which is

$$H_\phi = j\beta \frac{\mu I l}{4\pi r} \sin\theta e^{-j\beta r} \left(1 + \frac{1}{j\beta r} \right) \quad (34)$$

Following the procedure outlined earlier, we find the electric field at r from the Maxwell equation

$$\nabla \times \bar{H} = j\omega\epsilon\bar{E} \quad (35)$$

Use of 34 in 35 yields

$$E_r = \frac{\eta \cos\theta I l e^{-j\beta r}}{2\pi r^2} \left(1 + \frac{1}{j\beta r} \right) \quad (36)$$

$$E_\theta = j\beta \frac{\eta \sin\theta I l e^{-j\beta r}}{4\pi r} \left(1 + \frac{1}{j\beta r} - \frac{1}{\beta^2 r^2} \right) \quad (37)$$

$$E_\phi = 0 \quad (38)$$

Equations 34-38 simplify considerably at large distances from the antenna, i.e., when $\beta r \gg 1$. For this case the only surviving terms are

$$\begin{aligned} H_\phi &= \frac{\beta}{\mu} \frac{I l \sin\theta}{4\pi r} e^{-j\beta r} \\ &= j \frac{I l \sin\theta}{4\pi r} e^{-j\beta r} \end{aligned} \quad (39)$$

$$\begin{aligned}
 E_{\theta} &= j\beta \frac{\eta I l \sin\theta}{4\pi r} e^{-j\beta r} \\
 &= j \frac{\eta I l \sin\theta}{2\lambda r} e^{-j\beta r} .
 \end{aligned}
 \quad (40)$$

The terms given by 39 and 40 are the only significant fields when r is greater than 10 wavelengths or so and are called the far zone fields or radiation fields.

Near Zone, Quasi Statics

On the other hand when $\beta r \ll 1$, the equations reduce to

$$H_{\phi} = \frac{I l}{4\pi r^2} \sin\theta \quad (41)$$

$$E_{\theta} = -\frac{j\eta I l}{4\pi\beta r^3} \sin\theta \quad (42)$$

$$E_r = \frac{\eta I l}{j2\beta\pi r^3} \cos\theta . \quad (43)$$

One notes the absence of any phase delay in these equations, i.e., near to an antenna ($\beta r \ll 1$) the fields are quasi-static. Eq. 41 may be recognized as the field of a current element given by the Biot-Savart law for steady currents. Equations 42 and 43 can be rewritten in the form of the static electric fields of a charge dipole with the substitution from the continuity equation

$$I = j\omega Q .$$

Then 42 and 43 become

$$E_r = \frac{Ql \cos\theta}{2\pi\epsilon r^3} \quad (44)$$

$$E_{\theta} = \frac{Ql \sin\theta}{4\pi\epsilon r^3} \quad (45)$$

which along with 41 were developed in an elementary fields course.

Radiation Fields

We now return our consideration to the far zone fields, Eqs. 39 and 40. We may calculate the radiated power density by forming the Poynting vector

$$\begin{aligned}
 \bar{S}_{AVG} &= \frac{1}{2} \operatorname{Re} \bar{E} \times \bar{H}^* \\
 &= \frac{|I l|^2 \eta \sin^2\theta}{8\lambda^2 r^2} \hat{u}_r \quad \text{w/m}^2 .
 \end{aligned}
 \quad (46)$$

The power flow is in the $\hat{u}_r$ direction, radially away from the source. The fields are TEM to r and thus at large distances approximate a plane wave. In fact in limited regions at large r the fields approximate a uniform plane wave.

The radiation pattern of the antenna is a plot of the relative field intensity magnitude as a function of θ for fixed r . This plot, which is obtained from

Eq. 40, is sketched in Fig. 4. One notes the null fields at $\theta = 0, \pi$ along the axis of the current element and the radiation maxima at $\theta = \pi/2, 3\pi/2$ normal to the radiator. The power radiation pattern is just a polar plot of Eq. 46, i.e., for this case a plot of $\sin^2\theta$ instead of $\sin\theta$.

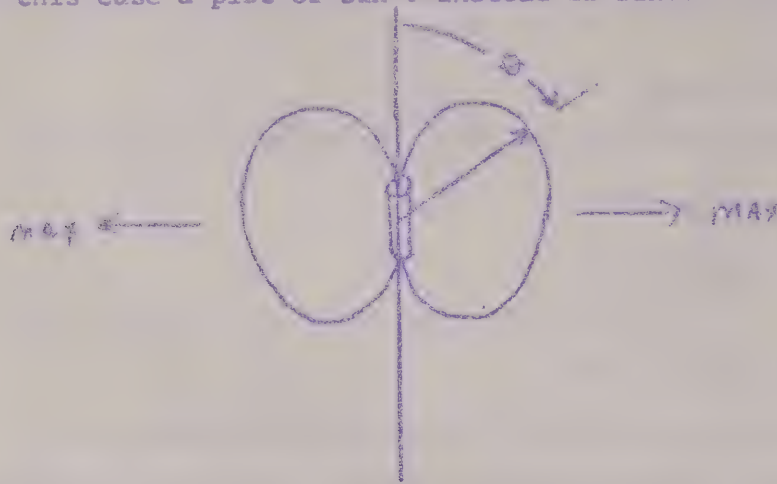


Fig. 4

The total radiated power is found by integrating Eq. 46 over the surface of a sphere enclosing the element. Thus

$$P = \int_0^\pi \int_0^{2\pi} \frac{|I_{\max} \ell|^2 \eta \sin^2\theta}{P 8\lambda^2 r^2} r^2 \sin\theta d\phi d\theta \text{ watts} \quad (47)$$

Note that this integration result is independent of r as would be expected from the conservation of energy. Carrying out the details of 47 results in

$$P_{\text{AVG}} = \frac{|I_{\text{rms}} \ell|^2 2\pi\eta}{32\pi} \text{ watts} \quad (48)$$

In free space where $\eta = 377 \Omega$, Eq. 48 yields

$$P_{\text{AVG}} = 789 I_{\text{rms}}^2 \left(\frac{\ell}{\lambda}\right)^2 \text{ watts} \quad (49)$$

From a circuits point of view we are delivering the average power of Eq. 48 to the terminals of the current element. Since the terminal rms current is I_{rms} , we could account for this power by an equivalent resistance at these terminals of

$$R_{\text{rad}} = 789 \left(\frac{\ell}{\lambda}\right)^2 \Omega \quad (50)$$

This resistance is called the radiation resistance and is simply the real part of the input impedance at the antenna terminals. The resistance given by 50 is an equivalent resistance which accounts for the radiated power and hence has nothing to do with dissipation in the conductor.

Radiation resistance per se is not a parameter to be considered in comparing antennas for relative merit. It is, however, most important as a system parameter since it is the load resistance that must be driven by the transmission line feeding the antenna. Efficiency, on the other hand, is determined by the radiation resistance in comparison with the loss resistance. The loss resistance for the current element must usually be determined from the surface resistance by the methods detailed earlier since skin effect will normally be a factor. This resistance will then be in series with the radiation resistance and the efficiency can be determined from

$$\eta = \frac{P_{\text{rad}}}{P_{\text{in}}} = \frac{I_{\text{rms}}^2 R_{\text{rad}}}{I_{\text{rms}}^2 (R_{\text{rad}} + R_{\text{diss}})} = \frac{R_{\text{rad}}}{R_{\text{rad}} + R_{\text{diss}}} \quad (50)$$

If R_{rad} is an order of magnitude greater than R_{diss} , further increase in R_{diss} will have no significant effect on the efficiency.

Half Wavelength Dipole

The function of an antenna is to couple electromagnetic energy into space. As such it serves as the coupling element between a bounded guiding system (e.g., transmission line) and unbounded space. From the driving line the antenna is viewed as a circuit element, i.e., a pair of terminals with an input impedance. From space it is viewed as an aperture, out of which or into which radiation passes. Both systems impose constraints on the design of the antenna. We have already seen that viewed from the driving system, radiation resistance is an important concept. In fact, the terminals of the current element present a complex impedance to the drive line. This might be expected since it was demonstrated that capacitive and inductive fields exist in the vicinity of the radiator. We have not calculated the input reactance but the student should be aware that it is a parameter that must be dealt with in designing a drive system. Viewing the antenna from space, the directional characteristics and field polarization are important parameters. The half wavelength dipole is a device that has useful directional characteristics and a real input impedance, and for these reasons it serves as a basic building block of many complicated antenna systems. Hence, we will analyze its behavior in some detail.

Figure 5 shows a center fed half wavelength dipole and the coordinate system used for the analysis.

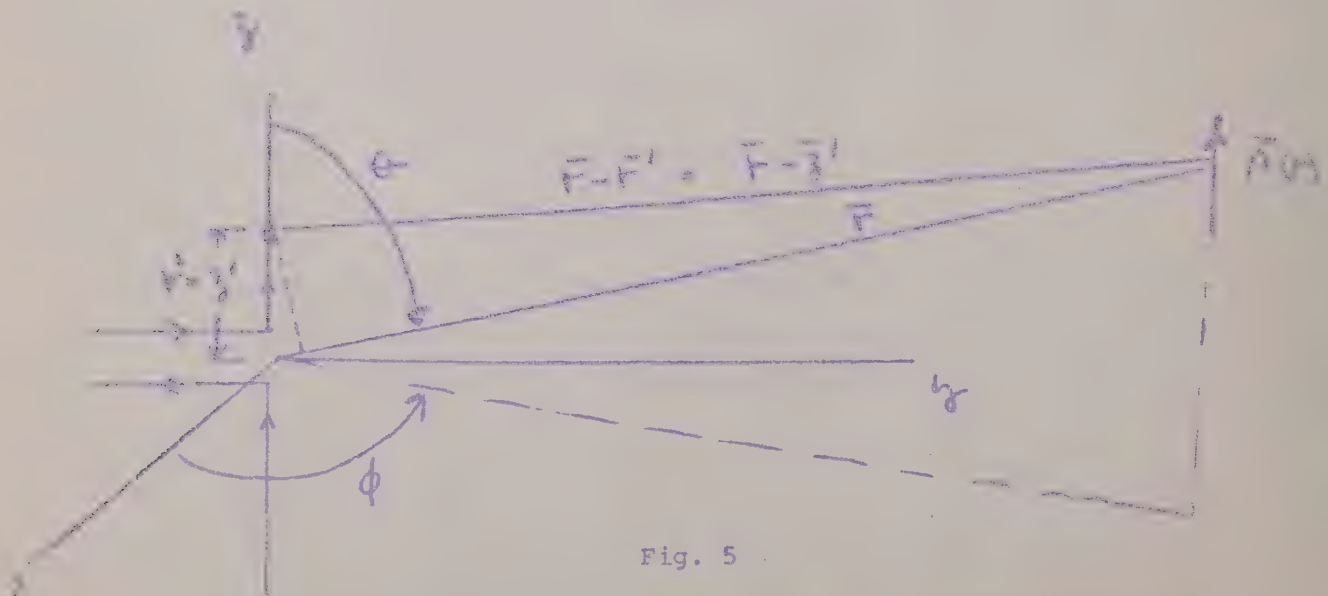


Fig. 5

Since the structure is long with respect to a wavelength we expect a variation in the current along the antenna. Because the ends of the antenna can be viewed as open circuits terminating transmission lines, the antenna currents will be standing waves and for a half wavelength structure will all

be in phase. The amplitude, however, will vary from zero at the ends to maximum at the center. This can be seen by considering the evolution of the dipole from a transmission line. In Fig. 6 as the open circuited line is bent into a dipole one sees the cosinusoidal standing wave develop. Since the model is

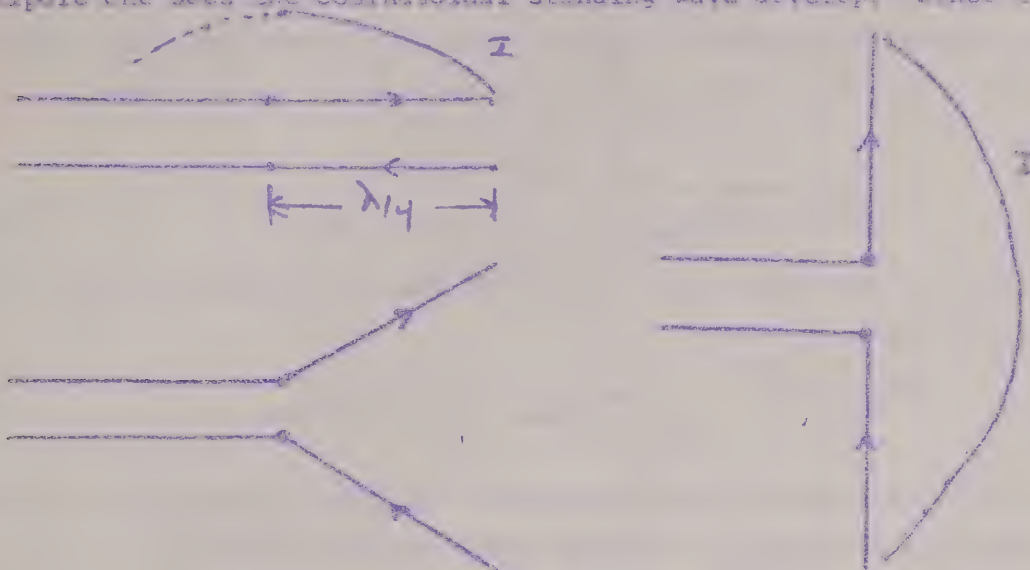


Fig. 6

over-simplified, the foregoing statements are not strictly true, but they are very nearly correct as is borne out by experimental measurement. Hence, for our analysis we will take the currents over the entire antenna to be in phase and vary cosinusoidally in amplitude about $z = 0$,

$$\vec{I}_c = I_{cmax} \cos \beta z \hat{u}_z \quad a/m^2 \quad (51)$$

We now consider the contribution to the vector potential at r of a current element of length dz at $r' = z'$. For such an element Eq. 23 becomes

$$\begin{aligned} d\vec{A}(r) &= \int \int_{\text{wire cross section}} \frac{\mu I_{cmax} \cos \beta z' e^{-j\beta |\vec{r}-\vec{z}'|}}{4\pi |\vec{r}-\vec{z}'|} \hat{u}_z dz' d\vec{r} \\ &= \frac{\mu I_{cmax} \cos \beta z' e^{-j\beta |\vec{r}-\vec{z}'|}}{4\pi |\vec{r}-\vec{z}'|} \hat{u}_z dz'. \end{aligned} \quad (52)$$

This differential contribution by an elemental length at z' must now be integrated over the entire length by integration to find the total effect at r . In order to accomplish this integration let us examine the term $|\vec{r}-\vec{z}'|$ by consideration of Fig. 5. If $r \gg z'$, that is, if we are in the far field, then

$$|\vec{r}-\vec{z}'| \approx r - z' \cos \theta \quad (53)$$

Furthermore, inspection of Eq. 52 reveals that $|\vec{r}-\vec{z}'|$ in the denominator is merely an amplitude factor and can be further approximated at large r by

$$|\vec{r}-\vec{z}'| \approx r \quad (54)$$

However, for the phase term in the numerator, the use of Eq. 53 gives

$$\beta |\vec{r} - \vec{z}'| \approx \beta r - \beta z' \cos \theta \quad (55)$$

We note that the second term on the right cannot be dropped in this case because it contributes a significant amount of phase shift as z' goes from $-\lambda/4$ to $\lambda/4$ regardless of the value of r . Thus, at large r , a reasonable approximation of Eq. 52 is

$$d\vec{A}(r) = \frac{\mu I_{\max} \cos \beta z' e^{-j\beta r} e^{+j\beta z' \cos \theta}}{4\pi r} \hat{u}_z dz' \quad (56)$$

Now to find the net contribution of all of the differential elements in the half wavelength antenna we form

$$\vec{A}(r) = \frac{\mu I_{\max} e^{-j\beta r}}{4\pi r} \hat{u}_z \int_{-\lambda/4}^{\lambda/4} \cos \beta z' e^{j\beta z' \cos \theta} dz' \quad (57)$$

The integral in 56 is a standard form (cf. Dwight, Tables of Integrals and Other Mathematical Data, p. 128, #577.2), and the result can be written as

$$A_z(r) = \frac{\mu I_{\max} e^{-j\beta r}}{2\pi \beta r} \left[\frac{\cos\left(\frac{\pi}{2} \cos \theta\right)}{\sin^2 \theta} \right] \quad (58)$$

We now proceed as we did earlier in the current element case in finding the electromagnetic fields. Expressing A_z in spherical coordinates by the use of Eq. 31 and then using Eq. 10 we find in the far field

$$E_\phi = j \frac{I_{\max} e^{-j\beta r}}{2\pi r} \left[\frac{\cos\left(\frac{\pi}{2} \cos \theta\right)}{\sin \theta} \right] \quad (59)$$

One should note that retaining any terms other than far field terms, i.e., $1/r$ dependence, would be invalid in the above since a far field approximation has already been used in obtaining Eq. 58. Now from Maxwell's equation (15) we obtain $\vec{H}$ as

$$E_\theta = \frac{j\eta I_{\max} e^{-j\beta r}}{2\pi r} \left[\frac{\cos\left(\frac{\pi}{2} \cos \theta\right)}{\sin \theta} \right] \quad (60)$$

Examination of Eqs. 59 and 60 indicates that E and H are in phase and that their ratio is η . Such would indeed be expected for these fields since they are essentially plane waves (TEM_z) far from the source. The bracketed terms in 59 and 60 are of course the radiation pattern factor for the half wave dipole, and this pattern is sketched in Fig. 7.

A considerable simplification in the procedure for finding the far field terms exists. One recalls from Eq. 13 that

$$\vec{E} = -\nabla \phi(t) - j\omega \vec{A}(t) \quad (61)$$

and by use of the Lorentz condition 18, Eq. 61 may be rewritten as

$$\vec{E} = \frac{\nabla(\nabla \cdot \vec{A})}{j\omega\epsilon} - j\omega \vec{A} \quad (62)$$

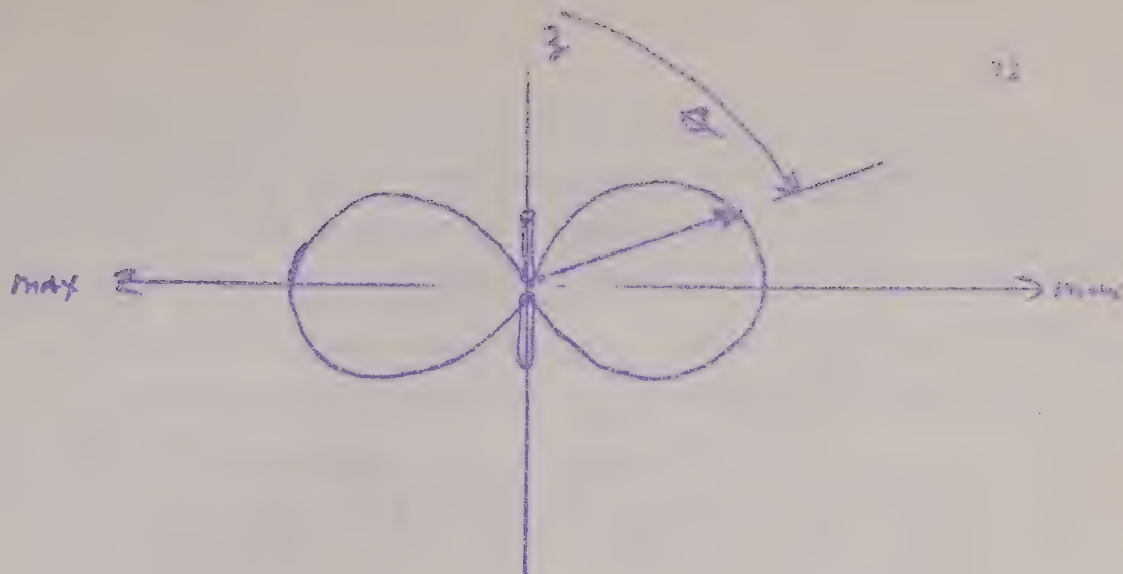


Fig. 7

thus allowing $\vec{E}$ to be obtained directly from a knowledge of $\vec{A}(t)$. At first appearance, Eq. 62 affords no simplification to the method already discussed. However, closer inspection of the first term reveals that it contributes nothing to the far field. This can be seen by inspecting the spherical coordinate operations of gradient and divergence as given, for example, in the text, p. 447, Eqs. D-32 and D-33. It is evident that for the transverse components (2,4) there will be only terms of higher order than $1/r$. Hence, we conclude that the second term in 62 is the only one of importance in the far field and,

$$\vec{E} = -j\omega \vec{A} \quad . \quad (63)$$

Hence, since from 31

$$A_\theta = -A_z \sin\theta$$

we may obtain Eq. 60 directly from 58. The magnetic field is also obtained with ease, since it is always true for plane waves in the far field that

$$E_\theta/H_\phi = \eta \quad . \quad (64)$$

We now proceed as before to determine the average power radiation from the Poynting theorem through

$$P_{AVG} = \frac{1}{2} \operatorname{Re} \oint (E_\theta \hat{u}_\theta \times H_\phi^* \hat{u}_\phi) \cdot d\vec{S} \quad . \quad (65)$$

Integrating 65 over the surface of a sphere of large radius using 33 and 36 results in

$$P_{AVG} = 73 I_{rms}^2 \text{ watts} \quad (66)$$

for a half wave dipole immersed in free space. We conclude, of course, that the radiation resistance seen at the center feed points is 73Ω .

Radiation resistance is not entirely a characteristic of a particular antenna. While the type of antenna places certain bounds on the radiation resistance, the driving point plays a much greater role in determining its value. For example, for the half wavelength dipole we have seen that the current is given by

$$\begin{aligned} I(z) &= I_{\max} \cos \beta z \\ &= I_{\text{loop}} \cos \beta z \end{aligned} \quad (67)$$

(The term I_{loop} is preferred over $I_{\max}$ to indicate the maximum value of the standing wave, thus avoiding confusion with the peak instantaneous value.) Now if we drive the antenna at some point, z , other than the center but maintain the maximum value of current, I_{loop} , the same, then the same power will be radiated. Thus

$$|I|_{\text{drive-rms}}^2 R_{\text{drive}} = |I|_{\text{loop-rms}}^2 R_{\text{center}} \quad (68)$$

Substituting from 67

$$R_{\text{drive}} = \frac{R_{\text{center}}}{\cos^2 \beta z} \quad (69)$$

where z is the driving point. One sees, for example, that the radiation resistance of a half wave dipole end fed, approaches infinity. In practice, resistances of the order of 1000 Ω are obtained.

The folded dipole shown in Fig. 8 is a common element in TV antennas. Such an antenna is just two dipoles in parallel, and hence one would expect

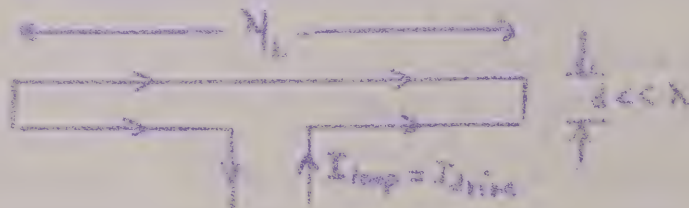


Fig. 8

that for the same maximum current, I_{loop} , the fields would be double. As a result the radiated power increases by a factor of four. Since the antenna is fed at a current loop

$$|I|_{\text{drive-rms}}^2 R_{\text{folded dipole}} = 4 |I|_{\text{drive-rms}}^2 R_{\text{dipole}} \quad (70)$$

or

$$R_{\text{folded dipole}} = 4(73) = 292 \Omega \quad (71)$$

Voilà, the famous 300 Ω TV antenna. It should not be implied that a folded dipole is 4 times more effective in radiating power; it just operates a higher impedance level than a dipole. Its choice for TV is dictated by its wider bandwidth not by its radiation resistance.

At this point it should be pointed out that the effective loss resistance seen at the driving point also depends on the choice of driving point. For example, for a dipole

$$P_{\text{dissipation loss}} = \frac{1}{2} \int_{-\lambda/4}^{\lambda/4} I^2(z) R_{ac} dz \quad (72)$$

where R_{ac} is skin effect dependent resistance parameter at the operating frequency. That is, for a wire antenna whose radius is large with respect to δ previous considerations yield

$$R_{ac} = \frac{R_s}{2\pi a} \Omega/m \quad (73)$$

where a is the wire radius and R_s is the surface resistance. With $I(z)$ given by 67 we find

$$\begin{aligned} P_{\text{diss}} &= \frac{1}{2} \int_{-\lambda/4}^{\lambda/4} I_{\text{loop}}^2 \cos^2 \beta z \frac{R_s}{2\pi a} dz \\ &= I_{\text{loop-rms}}^2 R_s \left(\frac{\lambda/4}{2\pi a} \right) \end{aligned} \quad (74)$$

One sees that because of the current amplitude variation, the half wave antenna dissipates as much power as a quarter wavelength wire with a uniform current.

The dissipated power can be written in terms of the driving point current for other than center drive by substituting from 67,

$$P_{\text{diss}} = \frac{I_{\text{drive-rms}}^2}{\cos^2 \beta z} R_s \left(\frac{\lambda/4}{2\pi a} \right) \quad (75)$$

Note that the dissipation loss resistance seen at the driving point is

$$R_{\text{diss-drive}} = \frac{R_s \left(\frac{\lambda/4}{2\pi a} \right)}{\cos^2 \beta z} \quad (76)$$

Using 69 the radiated power for an arbitrary drive point is

$$P_{\text{rad}} = \frac{I_{\text{drive-rms}}^2}{\cos^2 \beta z} R_{\text{center}} \quad (77)$$

Thus the antenna efficiency

$$\eta = \frac{P_{\text{rad}}}{P_{\text{rad}} + P_{\text{diss}}}$$

is independent of the drive point. For the half wave dipole η depends on the magnitude of $R_s \lambda / 8\pi a$ relative to 73 Ω .

Gain

An antenna is spoken of as having a gain because to an observer in the far field an antenna that is capable of concentrating power in his/her direction has exactly the same effect as increasing transmitter power to a non-directional antenna. In the fundamental definition of antenna gain in fact, the reference is taken as just such an isotropic radiator. For an isotropic or non-directional antenna, the power density at some radius r in the far field is

$$P_{iso} = \frac{P_{in}}{4\pi r^2} \text{ w/m}^2 \quad (78)$$

where P_{in} is the average driving point power in watts. For a directional antenna, the power density is given by

$$P(\theta) = \frac{|E_\theta|^2}{\eta} \quad (79)$$

where E_θ is the rms value of electric field intensity in the far field. Hence the antenna gain is

$$\begin{aligned} G(\theta) &= \frac{P(\theta)}{P_{iso}} \\ &= \frac{4\pi r^2 |E_\theta|^2}{\eta P_{in}} \end{aligned} \quad (80)$$

The maximum gain is the maximum directive gain or simply the directivity. It is the value usually quoted in handbooks, etc., as "gain". The maximum directive gain of a half wave dipole is obtained from 80 by use of 60 and 66 which gives

$$\begin{aligned} G_{\text{max dipole}} &= \frac{4\pi r^2 \eta^2 I_{rms}^2 \left[\cos\left(\frac{\pi}{2} \cos\frac{\pi}{2}\right) \right]^2}{4\pi^2 r^2 \eta I_{rms}^2} \\ &= 1.64 \end{aligned} \quad (81)$$

The identical calculation carried out for the current element using 43 and 49 yields

$$G_{\text{max element}} = 1.5 \quad (82)$$

By comparison of gain and directional characteristics as evidenced in Figs. 4 and 7, one might ask, why build half wavelength antennas when the current element behavior is so similar? This question is answered, of course, by viewing the antenna from the driving point. Even if the low efficiency could be accepted the problem of transforming the input impedance into a real value of reasonable level remains. Such transforming networks are most difficult

and if any amount of power is to be radiated, the low radiation resistance means high current and hence high loss. Thus the dipoles' advantages become apparent. Even so, short antennas do find application especially at low frequencies where long lengths (or heights) are impractical.

Certain loading techniques are used to cancel the reactive part of the driving point impedance of non-resonant length antennas. From experience with transmission lines one should expect for a short antenna that the input reactance would be capacitive. That is, it is an open circuited line less than a quarter wavelength long. Hence a series inductance at the terminals of such value to series resonate with this capacitance will provide a real driving point impedance. Such loading coils, in fact need not be placed at the terminals but may be connected in series along the antenna at a point of lower current in order to reduce the coil losses. This is shown in Fig. 9 along with capacitive loading at the open end. Such loading also produces a real input impedance and

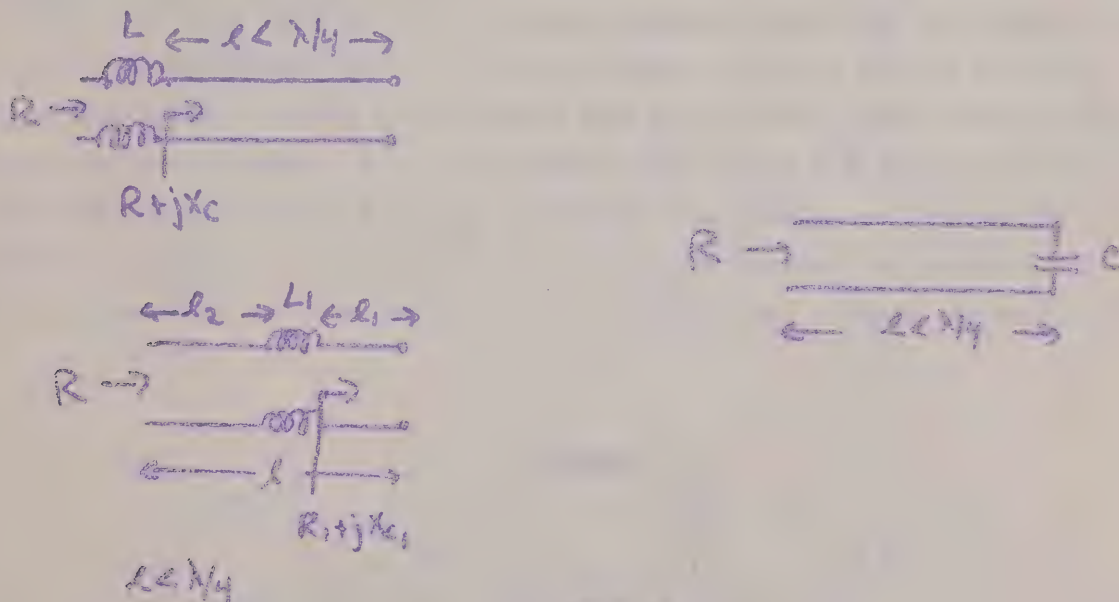


Fig. 9

antennas so constructed are said to be top loaded or loaded with capacitive hats. For the center loaded case, $l_1 < \lambda/4$ so X_{c1} is large. The series L_1 reduces X_{c1} to a small value, hence line l_2 sees a large C and transforms it in a short distance into a resistance. (Consider the Smith chart.)

One notes that the dipole antenna requires a balanced drive. If coax (unbalanced) line is to be used then a "balun" transformer discussed at the beginning of the semester must be used to provide the balanced currents. It is evident that if coax is connected directly to the dipole, radiating current

will appear on the outside of the coax sheath as well as on the antenna itself. The driving point impedance and of course the directional characteristics will be greatly altered.

Method of Images

To this point we have discussed only antennas immersed in free space. Since most antennas must radiate in the presence of conducting obstacles (e.g., the ground), let us see how this, at least in certain special cases, may be taken into account.

One will recall from electrostatics that image theory is based on the fact that conductors are equipotential surfaces and hence must have zero tangential E field. For the dynamic case we have similar conditions along with normal B equal to zero. A tangential H field will be terminated with a surface current. Imagine as shown in Fig. 10 a current element oriented parallel to a conducting plane of infinite extent. Radiated power is now reflected by the plane so we certainly expect the radiation pattern given earlier in Fig. 4 to be no longer valid. How then can we find the fields for $z > 0$? The uniqueness theorem, which will not be proved here, states that the fields which satisfy the boundary conditions on the surface are the unique fields for $z > 0$. Thus if we can construct an equivalent source to satisfy the condition $E_{\text{tan}} = 0$ on the metal, then this source will give us the desired fields in the region $z > 0$. Figure 11 shows an equivalent problem without the conductor at $z = 0$ but with an additional current

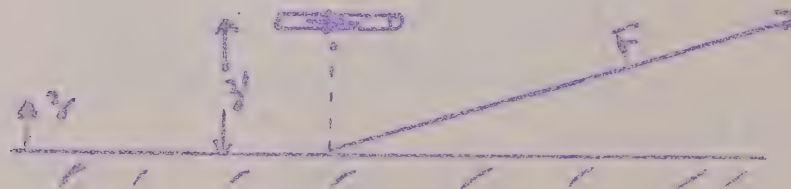


Fig. 10

element located at $-z'$ which together with the original element produces fields which are entirely normal to the surface at $z = 0$. Thus we conclude that the fields in the region $z > 0$ must be exactly the same as those in the example of Fig. 10. The second current element is called an image current and its fields in superposition with the fields of the original element are the desired fields in the region $z > 0$. The fields produced in the region $z < 0$ by the construct of Fig. 11 have no physical meaning and must be ignored. Note in Fig. 11 that as z' approaches zero the currents tend to cancel and radiation ceases. That

is, the current in Fig. 10 becomes shorted by the conductor. Such is just the case for a low frequency horizontal antenna too close to the ground.

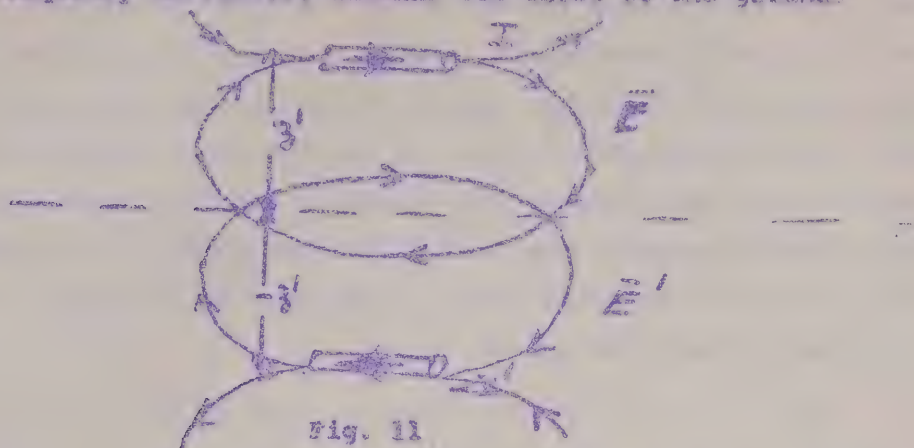


Fig. 11

Figure 12 shows the case for an element oriented normal to the surface. Here, the image current is in the same direction as the original. A current at an arbitrary angle may be considered as a combination of the currents and images of Figs. 11 and 12.

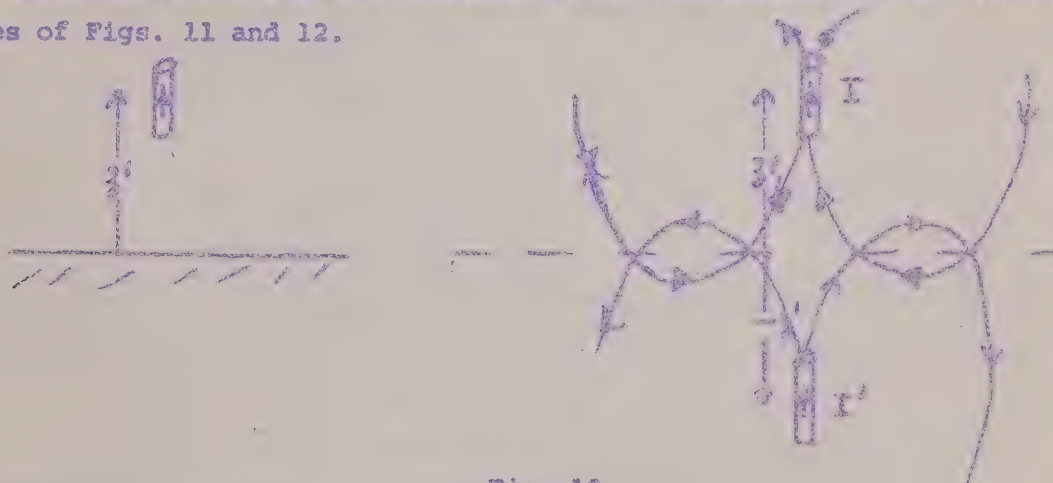


Fig. 12

The image idea may be applied to current distributions as well. For example, imagine a half wave dipole which has only normal E fields in the surface at $z = 0$ as shown in Fig. 13a. From the image concepts just presented, a

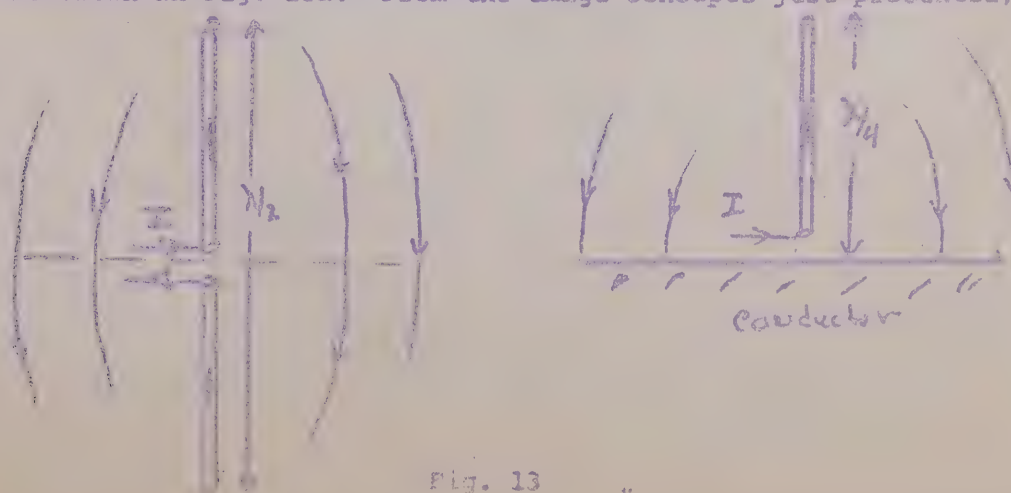


Fig. 13

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quarter wavelength antenna above a conducting plane (Fig. 13b) must produce exactly the same fields in the region $z > 0$ as the dipole. Such an antenna is called a quarter wave monopole and is a very practical antenna especially at low frequencies where the effects of the ground must be contended with. Since the terminal currents in Fig. 13 are the same, and the fields for $z > 0$ are the same but the monopole only radiates into half space, the radiation resistance of the monopole must be one half of that of the dipole. That is about 36Ω for the drive point shown. Note that the monopole is unbalanced and can be driven directly with coax.

Reciprocity

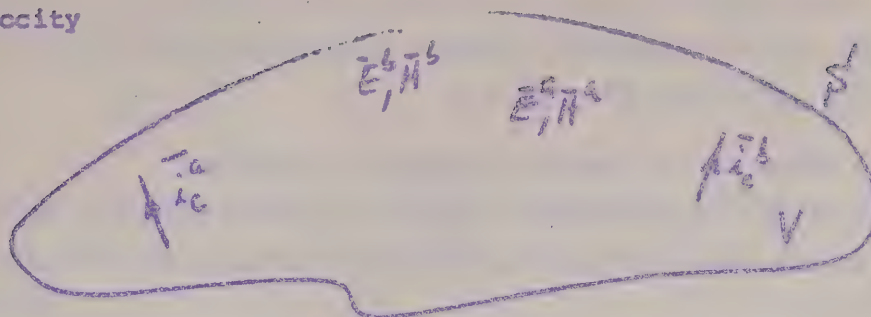


Fig. 14

The reciprocity theorem will be developed with the aid of Fig. 14. We consider two source currents enclosed within a homogeneous volume, \$V\$, bounded by the mathematical surface, \$S\$. The fields \$\vec{E}^a, \vec{H}^a\$ are produced by source current \$\vec{i}_c^a\$, and the \$\vec{E}^b, \vec{H}^b\$ fields are produced by the current \$\vec{i}_c^b\$. Hence, we relate the fields to the sources through Maxwell's equations

$$\nabla \times \vec{H}^a = j\omega\epsilon\vec{E}^a + \vec{i}_c^a \quad (83)$$

$$\nabla \times \vec{E}^a = -j\omega\mu\vec{H}^a \quad (84)$$

and

$$\nabla \times \vec{H}^b = j\omega\epsilon\vec{E}^b + \vec{i}_c^b \quad (85)$$

$$\nabla \times \vec{E}^b = -j\omega\mu\vec{H}^b \quad (86)$$

We now manipulate these equations by pre-dot-multiplying the first, 83, by \$\vec{E}^b\$, and pre-dot-multiplying the last, 86, by \$\vec{H}^a\$ with the result

$$\vec{E}^b \cdot \nabla \times \vec{H}^a = j\omega\epsilon\vec{E}^b \cdot \vec{E}^a + \vec{E}^b \cdot \vec{i}_c^a \quad (87)$$

$$\vec{H}^a \cdot \nabla \times \vec{E}^b = -j\omega\mu\vec{H}^a \cdot \vec{H}^b \quad (88)$$

By subtracting 88 from 87 and making use of the vector identity

$$\nabla \cdot (\vec{A} \times \vec{B}) = \vec{B} \cdot \nabla \times \vec{A} - \vec{A} \cdot \nabla \times \vec{B} \quad (89)$$

we can write

$$\nabla \cdot (\vec{H}^a \times \vec{E}^b) = j\omega\epsilon\vec{E}^b \cdot \vec{E}^a + j\omega\mu\vec{H}^a \cdot \vec{H}^b + \vec{E}^b \cdot \vec{i}_c^a \quad (90)$$

Applying the above procedure to Eqs. 84 and 85 yields an equation similar to 90, namely

$$\nabla \cdot (\vec{E}^a \times \vec{H}^b) = -j\omega\epsilon\vec{E}^a \cdot \vec{E}^b - j\omega\mu\vec{H}^a \cdot \vec{H}^b - \vec{E}^a \cdot \vec{i}_c^b \quad (91)$$

If now we add Eqs. 90 and 91 the final result of the manipulation is

$$\nabla \cdot (\vec{H}^a \times \vec{E}^b + \vec{E}^a \times \vec{H}^b) = \vec{E}^b \cdot \vec{i}_c^a - \vec{E}^a \cdot \vec{i}_c^b \quad (92)$$

If we again consider Fig. 14, the result of integrating Eq. 92 throughout the volume is, after applying the divergence theorem to the left hand side,

$$\oint_S (\vec{H}^a \times \vec{E}^b + \vec{E}^a \times \vec{H}^b) \cdot d\vec{S} = \iiint_V (\vec{E}^b \cdot \vec{i}_c^a - \vec{E}^a \cdot \vec{i}_c^b) dv \quad (93)$$

For a source free region in space which contains fields from two sources outside the volume, Eq. 93 predicts a special relationship on the boundary,

$$\oint_S (\vec{H}^a \times \vec{E}^b + \vec{E}^a \times \vec{H}^b) \cdot d\vec{S} = 0 \quad (94)$$

Equation 94 is known as the Lorentz reciprocity theorem.

For our present applications, Eq. 93 is a more useful form of the theorem. Consider the source currents in 93 and in Fig. 14 to be two antennas located in a volume in space.

Consider also that the boundary surface, S , is distant enough to be, at each point, in the far field of both antennas. Let us examine the left hand integral for this situation. In the far field of any z-directed antenna we know

$$E_\theta / H_\phi = \eta \quad (95)$$

hence, the integral becomes

$$\begin{aligned} \oint_S (\vec{H}^a \times \vec{E}^b + \vec{E}^a \times \vec{H}^b) \cdot d\vec{S} &= \oint_S E_\theta^a \hat{u}_\theta \times \frac{E_\theta^b}{\eta} \hat{u}_\phi - E_\theta^b \hat{u}_\theta \times \frac{E_\theta^a}{\eta} \hat{u}_\phi \cdot d\vec{S} \\ &= \left(\frac{E_\theta^a E_\theta^b - E_\theta^b E_\theta^a}{\eta} \right) r^2 \sin\theta d\theta d\phi \\ &= 0 \end{aligned} \quad (96)$$

We thus conclude

$$\iiint_V \vec{E}^b \cdot \vec{I}_c^a dv = \iiint_V \vec{E}^a \cdot \vec{I}_c^b dv \quad (97)$$

This form of the reciprocity theorem is commonly known as Rumsey's reaction theorem.

Suppose the two antennas above are half wave dipoles. The integrations of Eq. 97 are carried out with the aid of Fig. 15. Here an antenna is shown



Fig. 15

driven at the center with a current source. The volume of the Eq. 97 integrals is the antenna volume since $\vec{I}_c$ is zero elsewhere. Because E_{tang} is zero along the wire, these integrals reduce to

$$V_I^{b,a} = V_I^{a,b} \quad (98)$$

where V^b is the terminal voltage produced at antenna b by the fields of antenna a , and I^a is the total driving current of antenna a . A similar interpretation

holds for the right hand side. Rearranging Eq. 96 as

$$\frac{V^b}{I^b} = \frac{V^a}{I^a} \quad (99)$$

indicates the equality of transfer impedances. That is, the voltage at a due to b, V^b , divided by the current that produced it at b, I^b , is a transfer impedance.

From the point of view of circuit theory our problem is a two port, with antenna a source located at port 1 and antenna b source at port 2. Using circuit notation Eq. 99 becomes

$$\frac{V_{12}}{I_2} = \frac{V_{21}}{I_1} \quad (100)$$

The network associated with Eq. 100 is shown in Fig. 16. The describing equations for this network are

$$V_1 = I_1 Z_{11} + I_2 Z_{12} = V_{11} + V_{12} \quad (101)$$

$$V_2 = I_1 Z_{21} + I_2 Z_{22} = V_{21} + V_{22} \quad (102)$$



Fig. 16

The voltage V_{11} is the contribution to the total voltage V_1 of the source at port 1, I_1 , while the voltage V_{12} is the contribution of source I_2 at port 2. We see from 101 and 102 that

$$V_{12} = I_2 Z_{12} \quad \text{and} \quad V_{21} = I_1 Z_{21} \quad (103)$$

Using Eq. 100 in the above results in the equality of transfer impedances

$$Z_{12} = Z_{21} \quad (104)$$

As has been mentioned previously, circuit theory is just a specialization of field theory.

Equality of Transmitting and Receiving Patterns

Reciprocity may be used to show that an antenna behaves in an identical manner whether acting in a transmitting mode or in a receiving mode. The term, in fact, should be associated with the equipment attached to the antenna rather than to the antenna itself. Imagine an experiment as depicted in Fig. 17 in

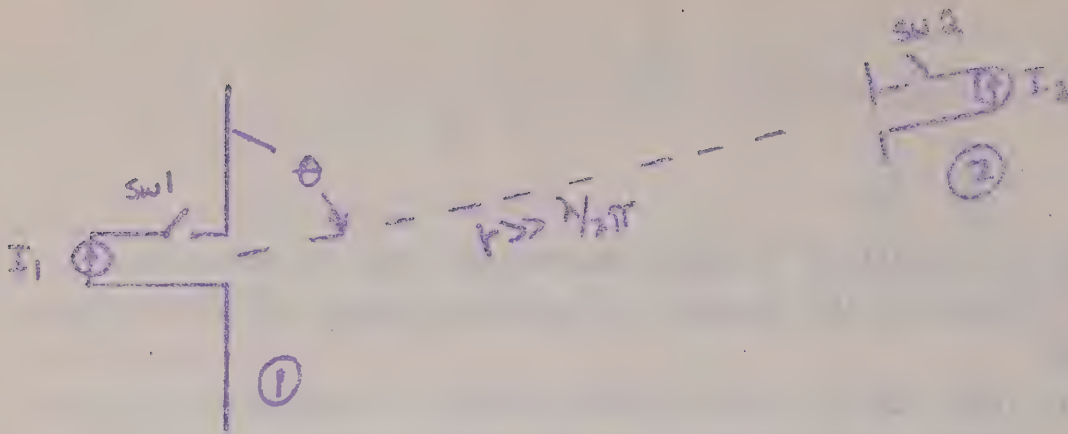


Fig. 17

which antenna 1 is an antenna for which the radiation pattern is known, for example the half wave dipole. Antenna 2 is an isotropic antenna located in the far field.

In experiment A we move antenna 2 on the surface of a sphere of constant radius and measure the received voltage at antenna 2 with switch 2 open and switch 1 closed. That is, we measure V_{21} , the voltage at 2 produced by the current at 1, and measure it as a function of θ , then

$$V_{21}(\theta) = KI_1 T(\theta) \text{ where } T(\theta) = \text{transmitting pattern factor and} \quad (105)$$

where K is a constant. We now perform experiment B in which switch 1 is opened, switch 2 closed, and the voltage at antenna 1 is measured as antenna 2 is again moved about on the sphere surface at constant radius. That is, we now measure $V_{12}(\theta)$, and from 100

$$\begin{aligned} V_{12}(\theta) &= \frac{I_2}{I_1} V_{21}(\theta) \\ &= KI_2 T(\theta) \end{aligned} \quad (106)$$

We see from 106 that the θ dependence for receiving is identical with that for transmitting.

It is interesting to note that the theorem of reciprocity predicts that electric current impressed upon perfect electric conductors cannot radiate. In Fig. 18 source a is the current on a conducting wire while source b is a line current in space. Equation 97 reveals that

$$0 = \iiint \vec{E}^a \cdot \vec{i}_c^b dv \quad (107)$$

The left hand side of 97 and hence 107 is zero because the boundary condition demands that $\vec{E}^b$, the field of source b, be zero at the conductor supporting current $\vec{i}_c^a$. Now since $\vec{i}_c^b$ definitely is not zero we must conclude that $\vec{E}^a$ is zero and hence source a produces no fields! In view of this startling fact,

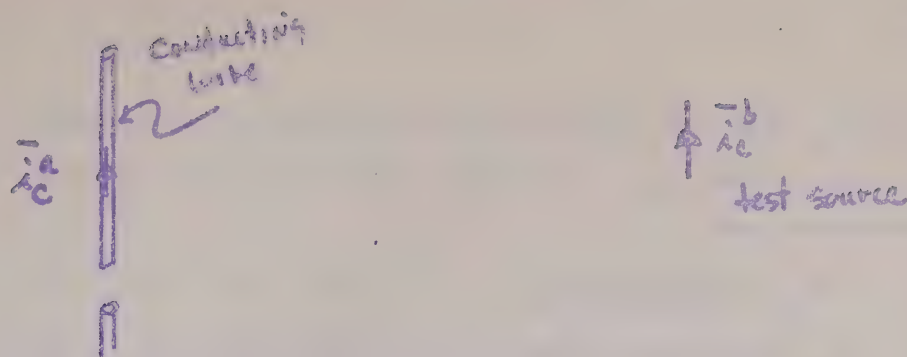


Fig. 18

is all of our previous work null and void? ----- No, because the antenna current never did radiate in the first place. The antenna is merely a reflector or scatterer of the radiant energy that we know from transmission line experience exists only in the gap at the feed point. This energy flows from the feed point into space and is directed into preferred regions by the presence of the reflecting conductors whose effective area as seen by the waves is much greater than the geometrical area.

Effective Area

The theorem of reciprocity will now be used to determine the effective area of an antenna. We consider here the maximum effective area, that is the effective aperture seen by a wave arriving at the antenna along the maximum direction of the radiation pattern which for a half wave dipole is $\theta = \pi/2$. Referring to Fig. 19(a) we invoke, by reciprocity, the equivalence of transfer impedance between any two pairs of terminals. Suppose one pair of terminals, 1, is at the drive point and the second set, 2, is taken across the element of length dz at z' . A voltage is induced along the antenna by the oncoming plane

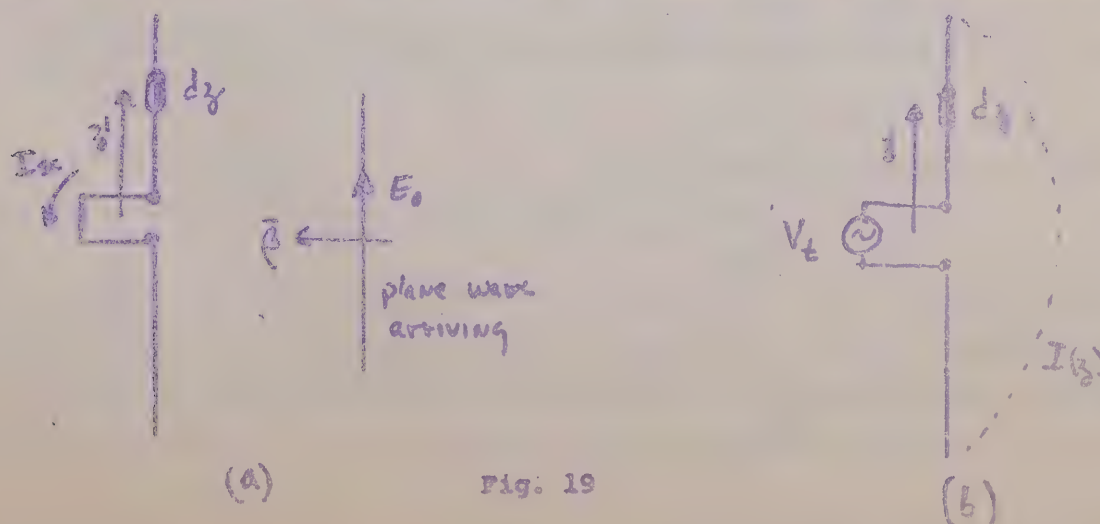


Fig. 19

wave field, E_0 , and its effect is seen at the short circuited drive point terminals as the current I_{sc} . A transfer impedance exists between the above-mentioned terminals and

$$Z_{21} = \frac{\text{voltage at 2}}{\text{current at 1}}$$

If we call dI_{sc} the contribution to the total current, I_{sc} , by the voltage at z' which is $E_0 dz$, then,

$$Z_{21} = \frac{E_0 dz}{dI_{sc}} \quad (108)$$

Consider now the same antenna connected to a voltage source, V_t , and operated as a transmitting antenna, Fig. 19(b). A standing wave of current appears on the antenna and is given by $I_{loop} \cos \beta z$. Now the transfer impedance can be written

$$\begin{aligned} Z_{12} &= \frac{\text{voltage at 1}}{\text{current at 2}} \\ &= \frac{V_t}{I_{loop} \cos \beta z'} \end{aligned} \quad (109)$$

By virtue of the reciprocity theorem, the impedances of Eqs. 108 and 109 are equal and

$$dI_{sc} = \frac{E_0 I_{loop} \cos \beta z' dz}{V_t} \quad (110)$$

The total short circuit current can now be found by integrating the differential contributions of Eq. 110 over the antenna length. Thus for Fig. 19(a)

$$I_{sc} = \frac{E_0 I_{loop}}{V_t} \int_{-\lambda/4}^{\lambda/4} \cos \beta z' dz' \quad (111)$$

The open circuit voltage available at the terminals in Fig. 19(a) with the short removed can be found from the Thevenin equivalent for the antenna and is

$$V_{oc} = I_{sc} Z_a \quad (112)$$

where Z_a is the driving point impedance. Since it is also true that,

$$Z_a = V_t / I_{loop} \quad (113)$$

we use 112 and 113 in 111 to give

$$V_{oc} = E_0 \int_{-\lambda/4}^{\lambda/4} \cos \beta z' dz' \quad (114)$$

The integral in 114 is defined as the effective length (or effective height) of the antenna. That is, the open circuit voltage available from the antenna is found by multiplying the incident uniform field intensity by the effective length. Note that because of the nonuniform current distribution on the antenna

the effective length is less than the actual length. For the half wave dipole, Eq. 114 yields

$$V_{oc} = \frac{E_0 \lambda}{\pi} \text{ volts} . \quad (115)$$

We now find finally the effective area of the antenna by equating the power available from the antenna to that delivered by the uniform plane wave. Maximum power transfer is obtained from a half wave dipole when a load equal to the conjugate of the driving point impedance is driven. For this case the maximum available power is

$$\begin{aligned} P_{rcvd} &= \frac{(V_{oc})^2}{4R_{rad}} \\ &= \frac{|E_0|^2 \lambda^2}{4\pi^2 R_{rad}} \text{ watts} . \end{aligned} \quad (116)$$

The power density in the arriving uniform plane wave is

$$P_{density} = |E_0|^2 / \eta \text{ w/m}^2 . \quad (117)$$

This power density multiplied ^{by} times the effective area is by definition the power of Eq. 116, hence

$$\begin{aligned} A_{eff} &= \frac{P_{rcvd}}{P_{density}} \\ &= \frac{\eta \lambda^2}{4\pi^2 R_{rad}} \text{ m}^2 . \end{aligned} \quad (118)$$

For a half wave dipole in free space

$$A_{eff} = 0.131 \lambda^2 \text{ m}^2 . \quad (119)$$

If we call the effective width of the antenna d_{eff} , then for the half wave dipole

$$d_{eff} = 0.262 \lambda$$

which is far greater than the geometrical width.

It is also interesting to note that the antenna gain calculation of Eq. 81 for a $\lambda/2$ dipole may be written as

$$G_{max} = \frac{\eta}{\pi R_{rad}} . \quad (120)$$

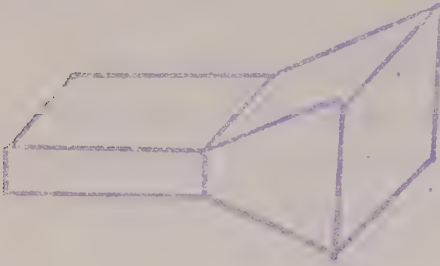
Taking the ratio of Eqs. 120 and 118 gives

$$\frac{G_{max}}{A_{eff}} = \frac{4\pi}{\lambda^2} . \quad (121)$$

We shall show shortly that this ratio is a constant for all antennas.

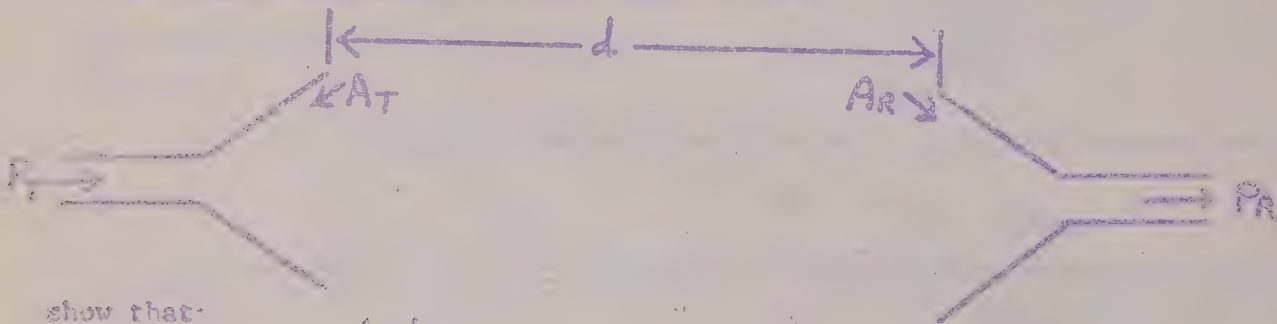
Free-Space Transmission Equation

The reason for presenting the derivation of the simple free-space transmission formula is to make clear a few of the concepts and definitions in antenna theory. Although microwave horn antennas will be used as an illustration, the results of the derivation will be quite general.



A horn antenna has an effective area both as a receiver or transmitter which is smaller than the geometrical area. In the equations to follow A will mean effective area.

Now for transmission in free space we can



show that:

$$P_R = P_T \frac{A_T A_R}{\lambda^2 d^2} \text{ watts}$$

P_T = power fed into transmitting antenna at its input terminals

P_R = power available at the output terminals of the receiving antenna

A_R = effective area of receiving antenna

A_T = effective area of transmitting antenna

d = distance between antenna apertures

λ = wavelength in free space

Now if P_T is the total power radiated and if it were to be radiated equally in all directions (isotropic radiator) then the power per unit area is

$$P_0 = \frac{P_T}{4\pi d^2} \text{ power flow/unit area}$$

$4\pi d^2$ = area of sphere

The directive power gain of an antenna is defined as

$$G_T = \frac{\text{Actual power per unit area in a given direction}}{\text{Power flow per unit area from isotropic radiator}}$$

$$G_T = \frac{P_a}{P_0} \text{ (Transmitting antenna)}$$

Thus

$$P_a = P_0 G_T = \frac{P_T G_T}{4\pi d^2} \text{ (power/unit area at receiver)}$$

Sometimes known as the "Friis Transmission Formula"

The effective area of an antenna is defined as

$$A_R = \frac{\text{Total power received}}{\text{Power per unit area at receiving antenna}}$$

$$A_R = \frac{P_R}{P_A} = \frac{P_R}{P_T G_T} = \frac{P_R 4\pi d^2}{P_T G_T}$$

or $P_R = \frac{P_T G_T A_R}{4\pi d^2}$ watts received power

In order to make clear the application of reciprocity at this point in the derivation of the free space transmission equation it will be convenient to temporarily enlarge upon the notation used thus far. Let

$$P_{RR} = \frac{P_{TT} G_T A_R}{4\pi d^2}$$

where P_{RR} = total power received at receiving antenna

and P_{TT} = total power transmitted by transmitting antenna

If now the transmitter (signal generator) is connected to the receiving antenna terminals and the receiver is connected to the transmitting antenna terminals, i.e. interchange the positions of the transmitter and receiver, then it is possible to write that

$$P_{RT} = \frac{P_{TR} G_R A_T}{4\pi d^2}$$

where P_{RT} = total power received at the original transmitting antenna

and P_{TR} = total power transmitted by the original receiving antenna.

Since the same transmitter (same power output) was used for both situations above $P_{TR} = P_{TT}$. When reciprocity is applied the conclusion is the $P_{RT} = P_{RR}$. Thus, the above two equations are equal to each other and result in

$$G_R A_T = G_T A_R$$

or

$$\frac{G_R}{A_R} = \frac{G_T}{A_T} = \frac{G}{A} = \frac{\text{Gain}}{\text{Area}} = \text{constant}$$

Thus G/A is a constant since no particular antenna was specified in the derivation. If we can calculate G/A for one antenna then it will be the same for all other antennas. The calculation is somewhat involved and will not be given here. For a dipole and hence for all other antennas

$$\frac{G}{A} = \frac{4\pi}{\lambda^2}$$

or let

$$G_T = \frac{4\pi}{\lambda^2} A_T$$

$$P_R = \frac{P_T G_T A_R}{4\pi d^2} = \frac{P_T \frac{4\pi}{\lambda^2} A_T A_R}{4\pi d^2}$$

refer to previous notes

Thus the free space transmission equation has been derived.

This equation assumes both antennas are matched and that no reflected wave is present. The distance between source and receiver must be

$$d \geq \frac{2D^2}{\lambda}$$

where D is the largest linear dimension of the largest of the two antennas used.

It is interesting to note that if the gain of an antenna can be measured by some means then the effective area can be calculated. Thus

$$A = \frac{G\lambda^2}{4\pi}$$

For example an isotropic or spherical radiator has a power gain of 1 and thus

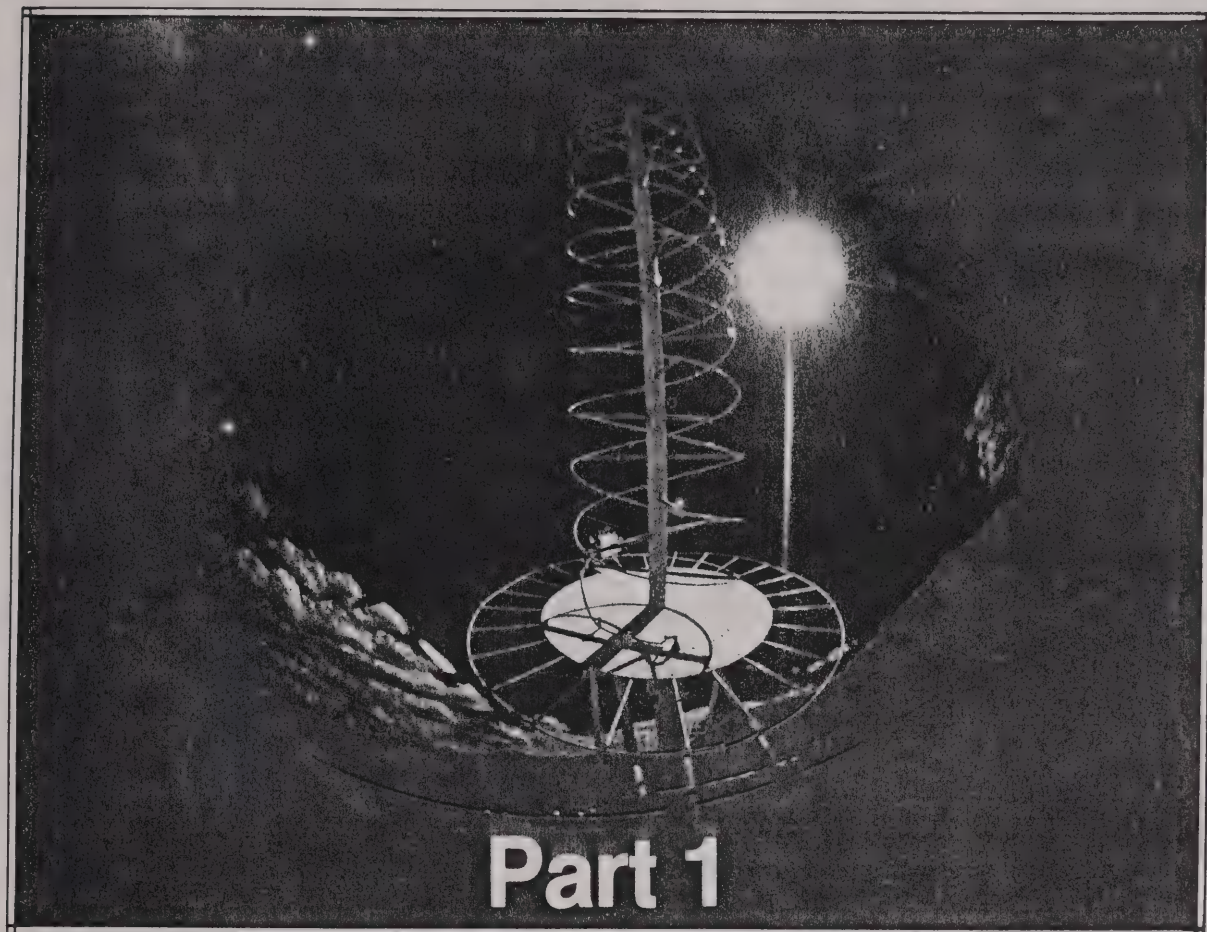
$$A(\text{isotropic}) = \frac{\lambda^2}{4\pi} \text{ square units}$$

Gains of other standard antennas are listed below:

	Power ratio	G(db)	A_{eff}
Isotropic (spherical)	1	0	$\lambda^2/4\pi$
Elemental dipole	3/2	1.76	$3\lambda^2/8\pi$
Circular (magnetic) dipole	3/2	1.76	$3\lambda^2/8\pi$
Half wave dipole	1.64	2.15	$0.13\lambda^2$

Antennas in the RF Range

... 10 kHz to 2000 MHz. A discussion of RF antennas used by all services.



Most antennas in the RF range fall into one of the following categories:

- A. Traveling wave
- B. Driven
- C. Vertical
- D. Leaky Waveguide
- E. Reflector

An antenna, as a transducer, converts electro-magnetic energy into RF currents for receiving applications or launches an electro-magnetic wave from an RF source(s).

Though there is the appearance of five separate categories there is some degree of overlap just as there is overlap in the frequency domain and usage of all antennas.

Vertical antennas, especially when combined in arrays are not really a distinct group. They fall into the traveling wave or driven antenna category. How-

ever, especially in the lower end of the RF spectrum (10 kHz-30 MHz), vertical antennas are widely used and are discussed as a separate group.

Leaky wave antennas, such as the Directional Discontinuity Ring Radiator, or DDDR for short, though they launch and receive a vertically polarized wave and consequently could be grouped as "C", due to their unique construction are considered separately.

Traveling Wave Antennas

Type 1. This is a group of endfire antennas utilizing an array of parasitic elements or wave guiding structure with energy directed towards or from a primary driven element. The categories are:

- 1. Yagi-Uda Parasitic Array
- 2. Backfire
- 3. Quad or closed loop array

4. Quagi (Quad-Yagi)
5. Helical
6. Birdcage

Once again not all categories are distinct. A back-fire antenna is a Yagi and reflector combination. However since modifications are required to the Yagi when combined with the reflective screen it is at best a hybrid and will be considered separately. (This is not meant to diminish the importance of it firing off the back and not the front end of the Yagi.)

The same holds true for the Quagi. Though constructed of "Yagi type" elements and closed loops by virtue of its hybrid performance rates a separate category.

1. Yagi-Uda Parasitic Array

The Yagi-Uda antenna, invented in 1926, is a parasitic structure that supports a slow surface wave along its length. It consists of a number of directors and reflectors (the parasitic elements) that enhance radiation in one direction.

The Yagi has been utilized by the military, commercial interest and experimenters alike. It is simple, rugged, can be made of all metal construction and consequently be DC grounded for lightning protection.

The simplest Yagi consists of a driven element and either a director or reflector. On the other extreme is an 80 wavelength long, 26 dBd (dB with respect to a reference dipole) experimental Yagi.¹ One of the longest boom Yagis in present use consists of 12 elements and is 152 feet long.¹⁰

A typical 3 element Yagi is depicted below.² (Figure 1)

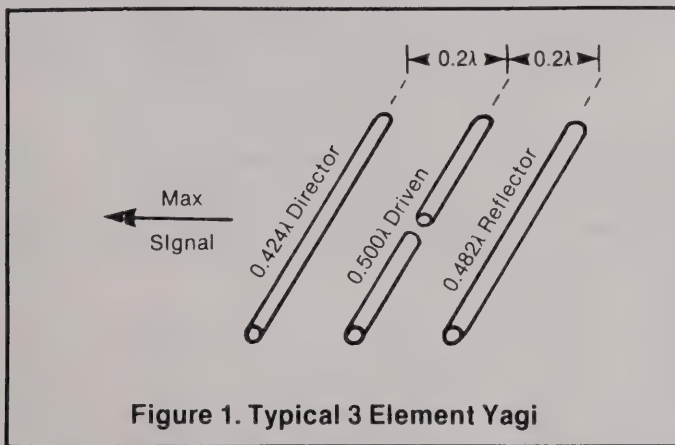


Figure 1. Typical 3 Element Yagi

Its elements can be linear, X-shaped or in the form of inverted Vees, each with its own merits.

a. Standard Design

There are several "recipes" for standardized construction of 3 or more element Yagis. One such recipe suggests that a single reflector be placed between 0.2 and 0.25 wavelengths behind the driven element and each added director of 2.5 percent diminishing length and 0.2 wavelength separation in front of the driven element.

The Hansen-Woodyard condition¹ specifies that high gains are achievable as long as the difference in

phase between the surface wave and the free space wave at the end of the antenna be equal to 180° . A general rule of thumb has always been "the longer it is the better it is (higher gain).

Recently several references have developed non-standard techniques for optimizing gain for a given boom length Yagi.

b. The Ascendency Curve Yagi³

Represents a design by formula called the ascendency — descendency curve. It recognizes the inter-relationships between the high number directors (5 and 6) and their effect in increasing the overall forward gain. Improvements of the order of 10 to 20 percent are indicated by the careful adjustments of non-uniform director *spacings*. An example of an 8 element, 13.1 dBd gain Yagi is given. For actual design information the reader is referred to the original source.³ (Figure 2).

D6 = 6th Director
DE = Driven Element
R = Reflector

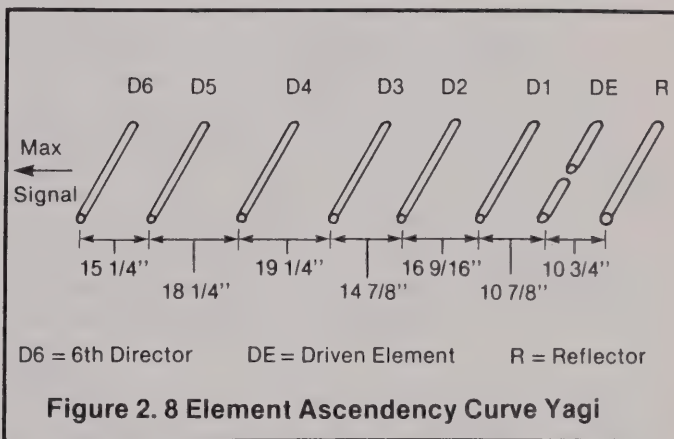


Figure 2. 8 Element Ascendency Curve Yagi

Note the ascending — descending spacings between the higher number directors. This Yagi is dimensioned in inches for operation at 222.5 MHz. (Which set of element lengths to use depends upon design bandwidth requirements and are clearly delineated in the reference.)

c. The Modulated Ladder Antenna

The modulated ladder antenna is basically a multi-element Yagi with non-uniform or "modulated" element lengths and spacings. The advantages indicated by this design are very high gain (26 dBd)¹ and minimum sidelobes over a substantial bandwidth. Properly designed modulated ladder arrays should exhibit an average beamwidth (H-plane beamwidth is larger, E-plane beamwidth is less) and gain as calculated from the formulas:

Beamwidth = $55\sqrt{l/\lambda}$ in degrees

Gain = $10 \log_{10}(9l/\lambda)$ in dB above isotropic

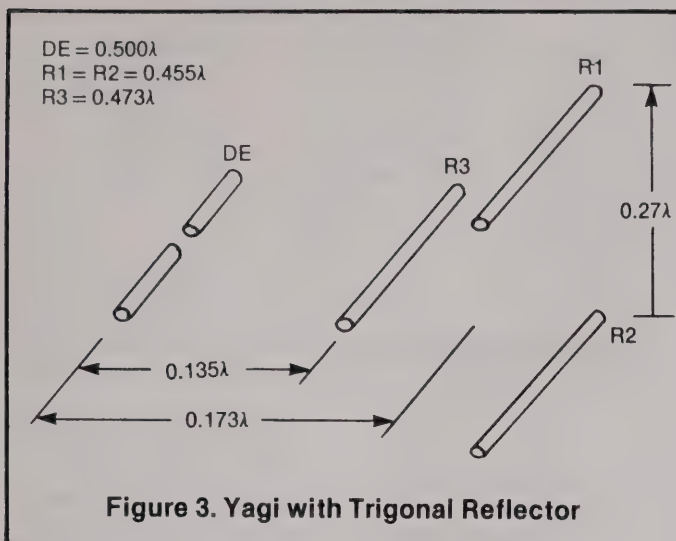
l = array length (same units as λ)

The concept relating higher achievable gains with non-uniform element lengths and spacings has been described in greater detail by David Cheng.^{4, 5}

d. The Trigonal Reflector

Most Yagi designs incorporate a single reflector. Experiments² utilizing plane reflecting surfaces, para-

bolas and corner reflectors were tried. The largest increase in gain (0.75 dB) occurred when a trigonal reflector was utilized as illustrated below in Figure 3.

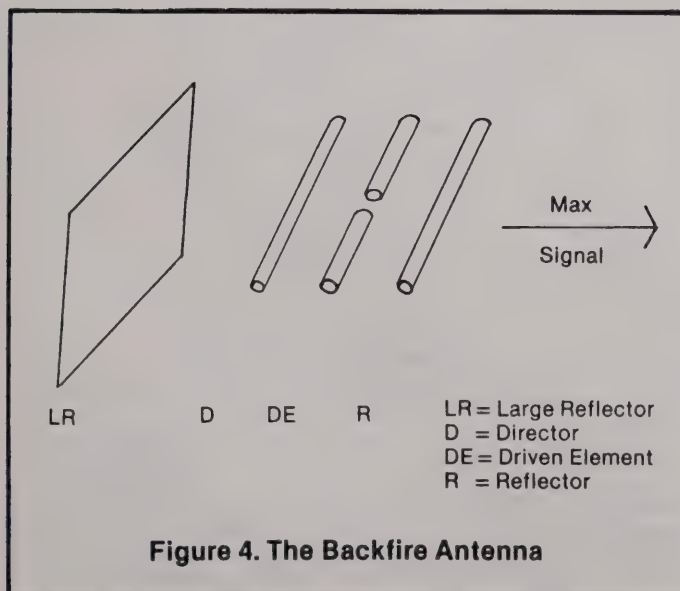


With well over 400 U.S. references to Yagi antenna development available since 1928⁶ and several definitive projects, some involving 3 or more years of research,² the Yagi antenna is probably one of the most interesting, useful and well studied antennas around.

2. Backfire Antenna

The Hansen-Woodyard condition, referred to previously, stated that the maximum gain of a Yagi is proportional to its length provided that the surface and free space wave 180° phase relationship is maintained. A Backfire antenna is basically a Yagi pointing towards a large reflector. (Figure 4)

LR = Large Reflector
D = Director
DE = Driven Element
R = Reflector



It takes advantage of the H-W condition by forcing the surface wave to traverse the physical length *twice*. This results in greater gain and frequency bandwidth than its equivalent length Yagi.⁷

Backfire antennas have been used at the high end of the RF range with gains in excess of 23 dB, at UHF with gains of 10.5 to 14 dB⁷ and at VHF with gain improvements of 4.5 dB over the isolated Yagi.⁸

3. Quad or Closed Loop Array

The cubical quad, or quad for short, was developed in 1942 by Clarence Moore.⁹ It is a parasitic traveling wave structure with loop elements approximately one wavelength in circumference. It is similar to the Yagi in several respects yet its differences are more widely known. These are:

- Approximately 2 dB greater gain than the equivalent length Yagi.
- "Quieter operation" — Less sensitive to precipitation static
- Feedpoint impedance closer in value to standard transmission line characteristic impedances
- Lower mounting height for the same takeoff or elevation angle as that of a Yagi.
- Polarization can be changed by simply changing feedpoints.

Handbook style design criteria for a 3 element quad follows from these basic formula:

- Circumference of driven element = 1005/fMHz feet
 - Circumference of reflector element = 1030/fMHz feet
 - Circumference of director element = 975/fMHz feet
- Inter-element spacings of from 0.15 λ to 0.20 λ are typical.

Power gain in dB above an isotropic of a closed loop is more a function of its perimeter and not shape. This has given use to loop designs with other than a diamond or rhombus configuration.

• Circular Loops

Research on large circular loop arrays¹¹ have indicated that the loop perimeter (or enclosed area) and relative thickness of the conductors are gain determining factors. Antenna area and gain are related as follows:

$$G = 4 \pi A / \lambda^2$$

where A (area) and λ² are entered with the same units of measure.

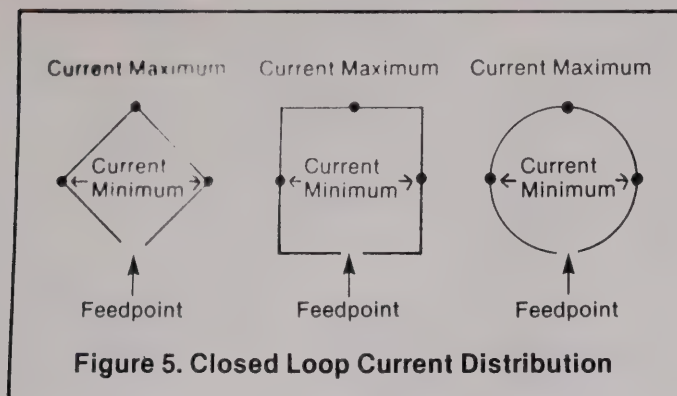
Figure 5 illustrates current distribution in diamond, square and circular 1 wavelength circumference loops:

• Delta Loops

Another popular form of loop array has a delta or triangular shaped element. It has the added advantage, especially at the lower end of the HF spectrum, in that it only requires a single high support, per element.

Delta loops exhibit substantially different take-off angles when inverted versus non-inverted, fed

at a corner or side and elevated versus erected close to the ground or ground system.^{12 13 14}

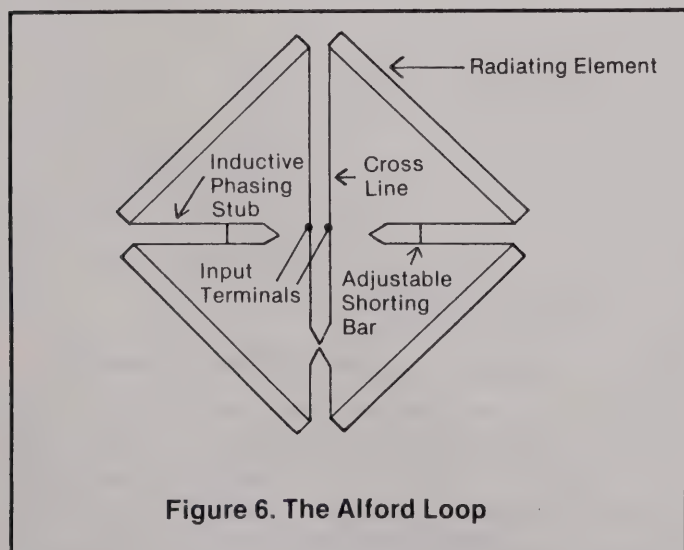


● Bi-square Array

As mentioned earlier conventional cubical quads have a 1 wavelength perimeter. A diamond-shaped quad would then have 4 sides each $\lambda/4$ long. The Bi-square loop is similar in shape and radiation direction (broadside) to the quad with the added advantage of 3 dB more gain. This additional gain is achieved by making the sides $\lambda/2$ in length. The total perimeter of a Bi-square loop is $4 \times \lambda/2$ or 2λ . By adding a parasitic reflector or director gains over 9 dBd have been claimed.¹⁵

● Alford Loop

One means of increasing antenna efficiency is to increase its radiation resistance. The Alford loop is a large square loop, consisting of four dipole radiators for sides. The radiation resistance is much higher than that of an ordinary loop. A typical Alford loop¹⁶ is indicated in Figure 6.

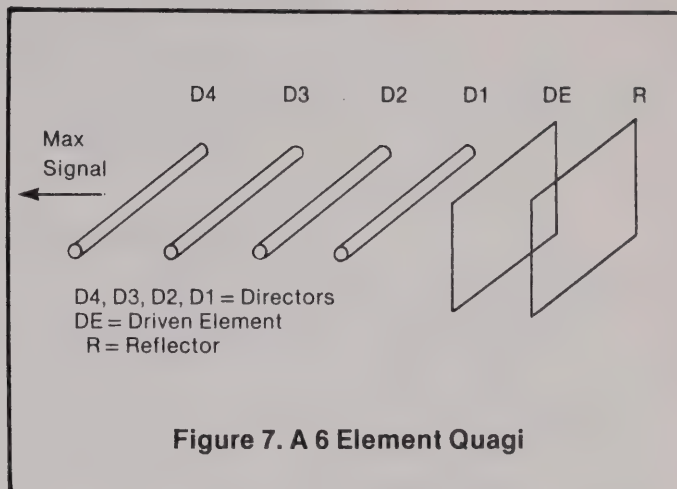


4. Quagis

The Quagi is a hybrid antenna that combines the best features of the cubical quad and the Yagi beam. Its directors are linear elements as in a Yagi while the driven and reflector are closed loops.

The immediately apparent advantages are greater gain over an equivalent number of element Yagi and simpler and more repeatable construction techniques. (Figure 7)

D4, D3, D2, D1 = Directors
DE = Driven Element
R = Reflector



5. Helicals

The front cover of this month's issue depicts a right-hand circularly polarized helical antenna operating at 468 MHz. Unlike dipole antennas where the electro-magnetic wave travels at the velocity of light the velocity of the wave along the axis of the helix is lower. It depends on the frequency, diameter and number of turns per unit length.

Helical antennas are categorized by their shape (cylindrical, flat or conical) and their mode of operation: normal or axial.

● Normal Mode

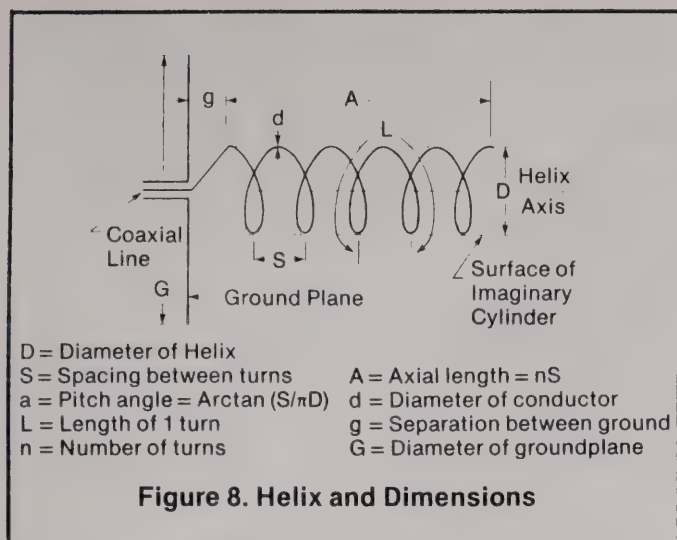
Distinguishing characteristics of a normal mode helix are their small diameters and lengths with respect to wavelength. The radiated field is elliptically polarized. Circular polarization is obtained with a resonant helix at a 0.9 height/diameter ratio.¹⁷

● Axial Mode

The axial mode helix is used as an end-fire circularly-polarized radiating beam. Gain is nearly constant over an octave. Both gain and beamwidth are a function of the number of turns — gain increases, the beamwidth decreases with increasing turns.

Axial mode helices are used for communications between terrestrial stations and orbiting satellites. Figure 8 illustrates a helix and its dimensions utilized for just such an application — radio amateur communications with OSCAR 7.²⁶

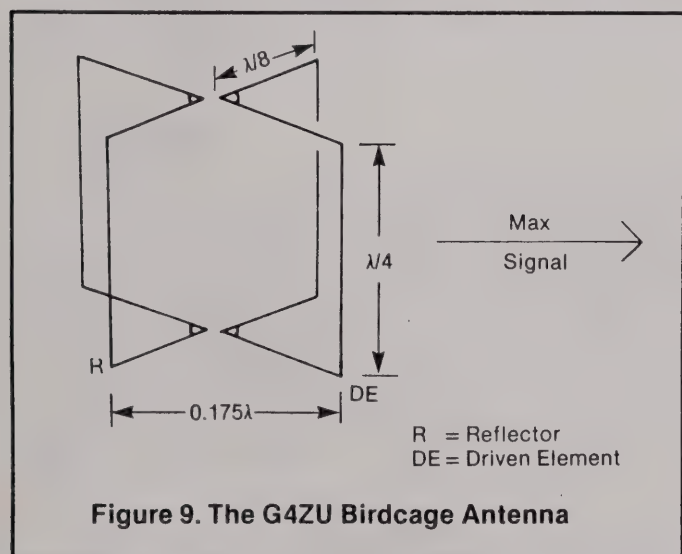
D = Diameter of Helix
S = Spacing between turns
 $a = \text{Pitch angle} = \text{Arctan}(S/\pi D)$
L = Length of 1 turn
n = Number of turns
A = Axial length = nS



d = Diameter of conductor
 g = separation between ground plane and first turn
 G = Diameter of ground plane

6. Birdcage

The birdcage is a unique 2 element parasitic array that borrows favorable characteristics from the cubical quad and Vee beams. Invented and patented by Dick Bird, a British radio amateur, in 1958,¹⁸ it closely resembles a bird cage. (Figure 9)



Several attractive features of the Birdcage are:

- High gain — between 8.5 to 10 dBd.
- Extremely compact — very small turning radius even at the low end of the HF spectrum.
- Feedpoint impedance — 40 to 50 ohms, ideal for matching to standard coaxial transmission lines.

Traveling Wave Antennas

Type 2. This is a group of antennas using a wave guiding structure, primarily wire, that is long

with respect to the operating wavelength. These antennas are additionally grouped by their height above ground.

• Elevated above ground ($>\lambda/8$)

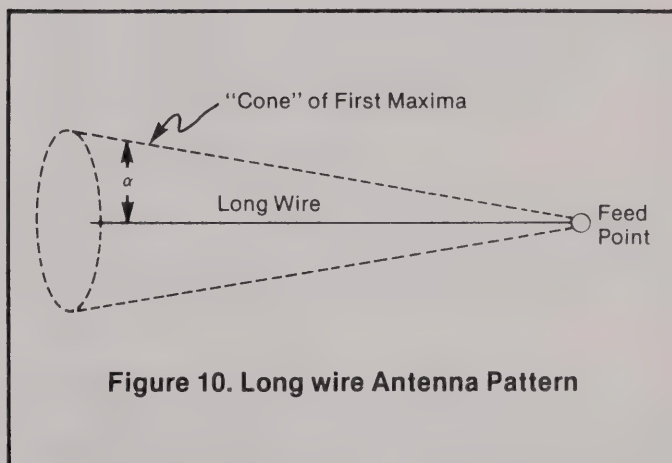
- Long wire antennas
- Vee antennas, Vee beams
- Rhombics

Unlike the previous Type 1 (Traveling Wave Antenna) Type 2 antennas evolve. That is long wires "1" make up Vee antennas "2" and Vees combine to form Rhombics "3".

1. Long-wire Antennas

The old adage "the bigger the better" applies to this type of antenna. A long wire antenna when fed at one end either directly (Fuchs antenna) or by means of a transmission line has a radiation pattern that directly relates to its length and height above ground. Radiation from end-fed long wire antennas is broadside when the overall length is $5/8$ wave or less.

As the length of the long wire increases above $5/8$ wave the radiation pattern forms into a cone coaxial and away from the feed point as illustrated below. (Figure 10)



A table of the locus (angle) of first maxima of the cone of radiation versus long wire length is given below: (Table 1)

Table 1			
α	λ	α	λ
53°	1	23°	5
37°	2	20°	6
29°	3	18°	7
26°	4	17°	8

The advantages of this type of antenna are its simplicity, high gain (for many wavelength long wires) and low cost to install and maintain.

2. Vee Antennas or Vee Beams

Vee antennas represent a natural progression from long wire antennas. They have the added ad-

vantages of greater gain, (3 dB versus same length long wires) sharper directivity and reduction of some sidelobes. Two long wires of specified length (λ) can be combined at one point to form a Vee beam of related angle (2α). This is to say that length and apex angle are related in optimized gain Vees.

Figure 11 illustrates spatial lobe combination and

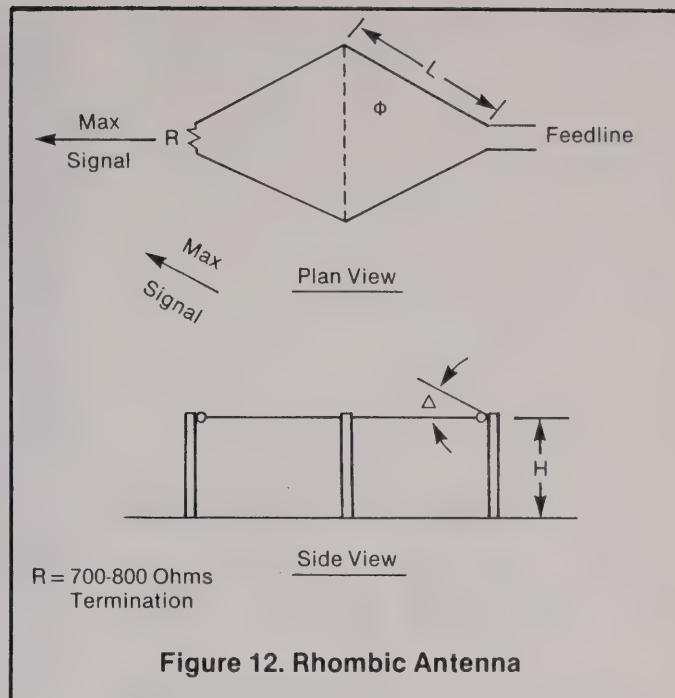
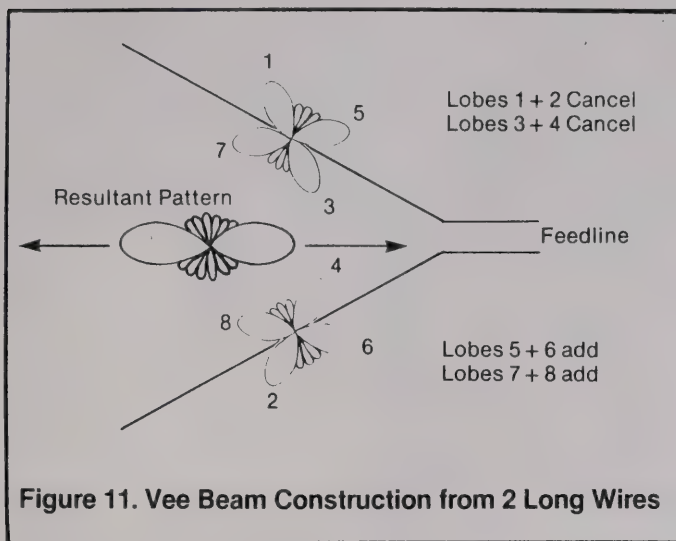


Figure 12. Rhombic Antenna

cancellation in the long wire construction of a Vee beam.

As illustrated a Vee beam, when properly constructed is bidirectional with maximum radiation along the bisector.

When resistively terminated at the ends the Vee Beam is unidirectional and radiates towards (and receives from) the left.

3. Rhombics

a. General

A rhombic is a four-sided, diamond-shaped high gain wire antenna. It provides a high signal-to-noise ratio for reception and a relatively low take-off or elevation angle. It exhibits a nearly constant feedpoint impedance over a wide bandwidth when properly terminated at the far end. A terminated rhombic is illustrated below. (Figure 12)

Maximum radiation is indicated along the bisector towards the left. Design terminology and criteria are as follows:

- ϕ = tilt angle
- Δ = vertical angle
- H = height above ground
- L = long side length

Using readily available charts¹⁹ and assigning values to two out of four of the parameters the remaining two are calculated.

Assume a 15° vertical angle is required. (Based on a specific propagation path and mode) and a maximum side length of 5λ is all the existing property will allow.

What tilt angle and height above ground does that calculate out to?

The referenced chart indicates that a tilt angle ϕ of 71° and a height of 0.95λ or 2.8λ will suffice. The lower of the two (heights) for practical considerations is chosen.

Terminations for rhombics are non-inductive resistors or in the case of very high power transmitters lossy balanced transmission line is utilized.

b. Multistrand Rhombics

Often for broadbanding purposes or for lowering the feedpoint impedance rhombics are multistranded. That is each leg consists of two or more conductors separated by several feet at the side poles and brought together at the forward and rear supports.

c. Nested Rhombics

A nested rhombic configuration is a pair of staggered coplanar rhombics with a common feedpoint. It exhibits substantial gain (27 dBd), broadbandedness, low vertical angle and lower (sidelobes) than a conventional rhombic.

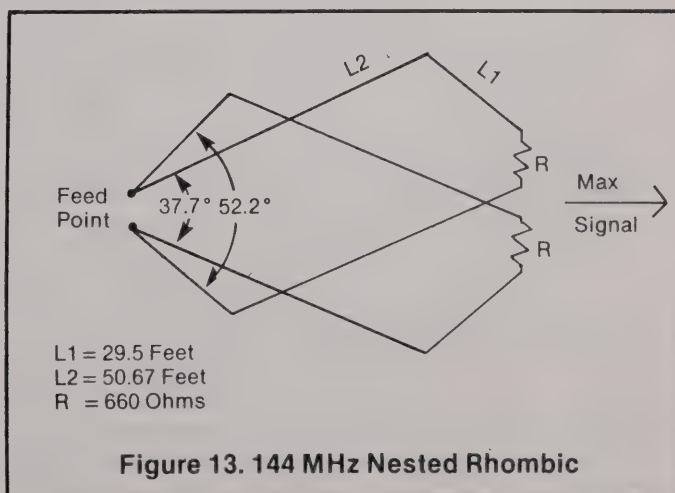


Figure 13. 144 MHz Nested Rhombic

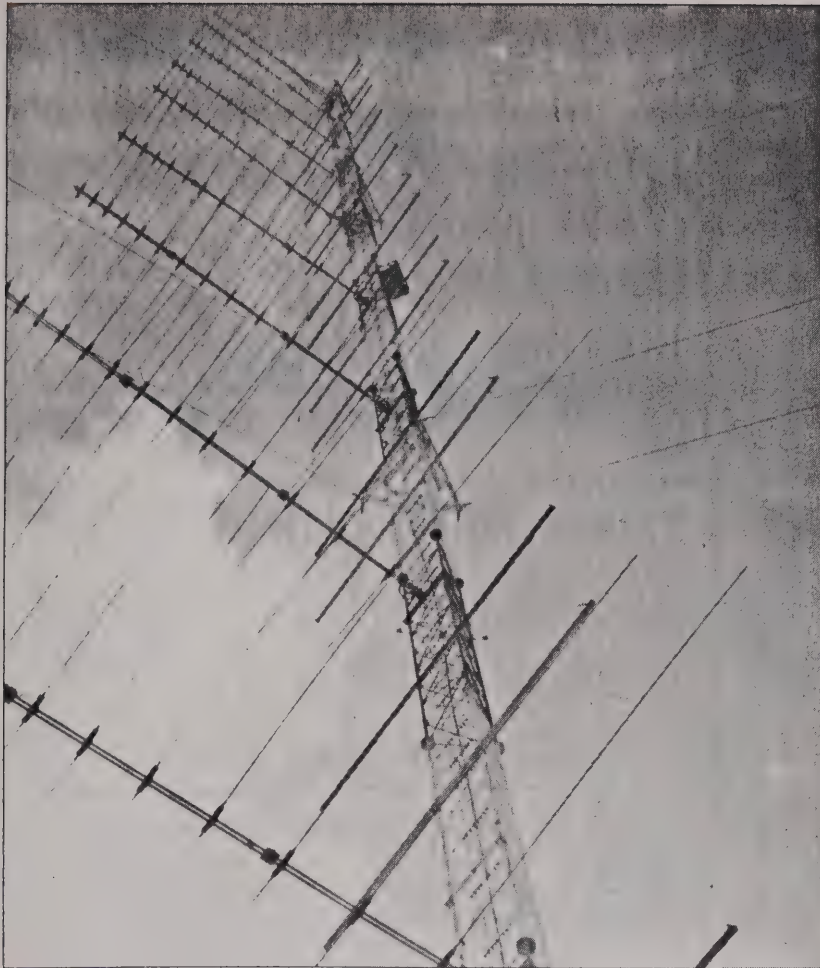


Photo of log periodic stacked array used in Artificial Radio Aurora experiments courtesy of Mr. Russell B. Stoner, N.T.I.A., Boulder, Colo.

Antennas In The RF Range

... 10 kHz to 2000 MHz. The discussion of RF antennas concludes with Driven, Vertical, Leaky-Wave and Reflector types considered.

Driven Antennas

A driven antenna can be defined as an array of two or more elements, separated in phase (electrical and/or spatial), hard connected via conductor or transmission line(s) and combined so as to develop a specific azimuthal or elevation pattern.

Driven antennas can be "loosely" subdivided into *Broadside* and *Endfire* categories. "Loosely" is meant to imply that some Broadside antennas when spatially rotated 90° become Endfire arrays and vice versa.

● Broadside

A broadside configuration is characterized by radiation (or reception) perpendicular, or steerable from the normal or perpendicular to the plane of the antenna or array. Four distinct types of Broadside arrays are considered though it must be reiterated that certain types of Endfire arrays can by spatial reorientation be considered Broadside.

The four types are:

1. Curtains
2. Dipoles, Collinears
3. Trombone T
4. Lazy H

Curtains

a. General

A curtain consists of an array of identical radiating elements configured in the vertical plane and in some cases backed by a reflecting screen.

Narrow beams and high broadside directivity gains are achieved by controlled amplitude and phasing techniques. Linear phase variations across the array enable the steering of transmitted or received wavefronts.

A uniformly spaced, uniform amplitude array has the advantage of producing the narrowest azimuthal beamwidth.

A uniformly spaced, tapered amplitude array such as a Binomial or Dolph-Tchebyshev distribution is utilized to minimize sidelobe level — a very useful technique in reducing susceptibility of a receiving antenna to interference.

b. Specific Designs

Each of the curtains illustrated below have substantial gain, various feedpoint locations and are inexpensive to build.

(1.) Sterba Curtain (Figure 15)

Probably the best known broadside array made up of a combination of parallel and collinear elements is the Sterba curtain.

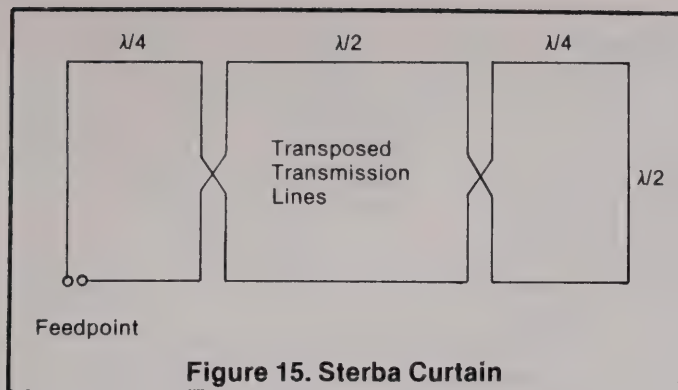


Figure 15. Sterba Curtain

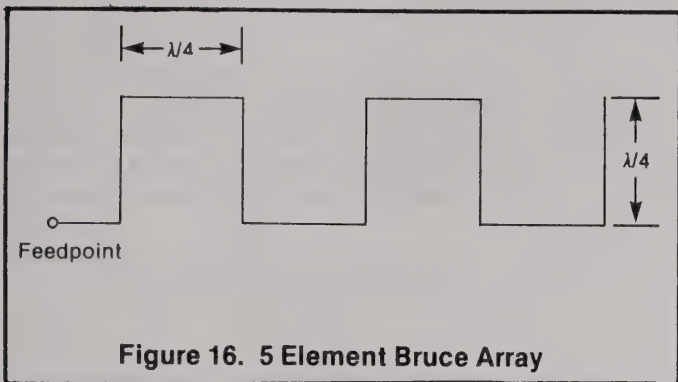


Figure 16. 5 Element Bruce Array

(2.) 5 Element Bruce Array (Figure 16)

The Bruce Array is configured with alternating $\lambda/4$ long vertical and horizontal elements. The radiation from the vertical sections are in phase and add while the radiation from the horizontal sections are out of phase and cancel. Typical gain of the 5 element array is 4 dB above a single quarter-wave vertical element.

(3.) The Chireix-Mesny Curtain

The C-M curtain consists of $\lambda/2$ dipoles configured in the form of diamonds. It is a center fed array with a broadside vertically-polarized radiation pattern. (Figure 17)

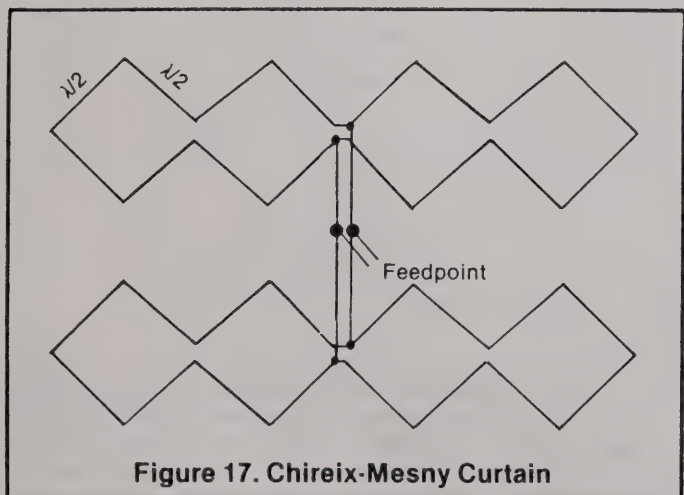


Figure 17. Chireix-Mesny Curtain

(4.) Bobtail Curtain

The Bobtail Curtain is a broadside array of three top current fed quarter wave verticals. Each vertical element section is separated from the middle section by 180° . By virtue of only feeding the center vertical a

1:2:1 amplitude distribution occurs with its associated negligible sidelobe pattern. Theoretical gain is 4.9 dB over a single $\lambda/4$ vertical (Figure 18)

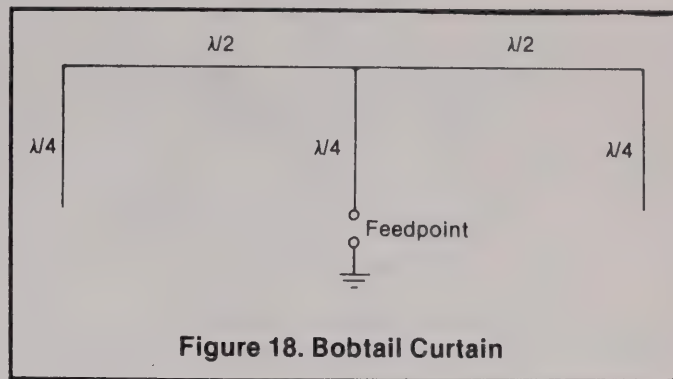


Figure 18. Bobtail Curtain

(5.) X-Array

The capital X array as indicated in Figure 19 has a power gain of 7 dBd.

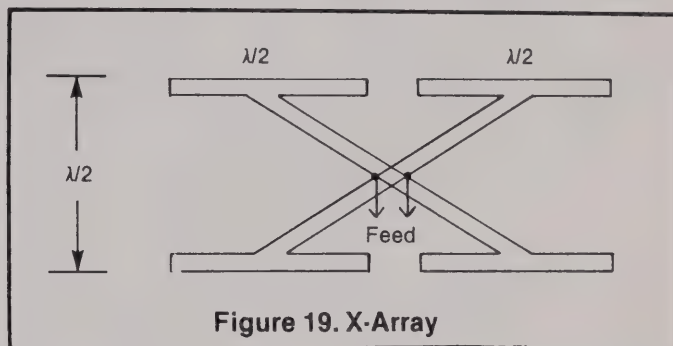


Figure 19. X-Array

Dipoles and Collinears

a. Unstuffed

There are two distinct categories of unstuffed dipoles: center and off-center fed.

(1.) Center-fed Dipoles

Flat-tops and inverted-vees. Both configurations and important properties are tabulated below by length:

Length	Name	Gain
$\lambda/2$	half-wave	0 dBd*
λ	double Zepp	1.8 dBd
1.28λ	extended double Zepp	3.0 dBd

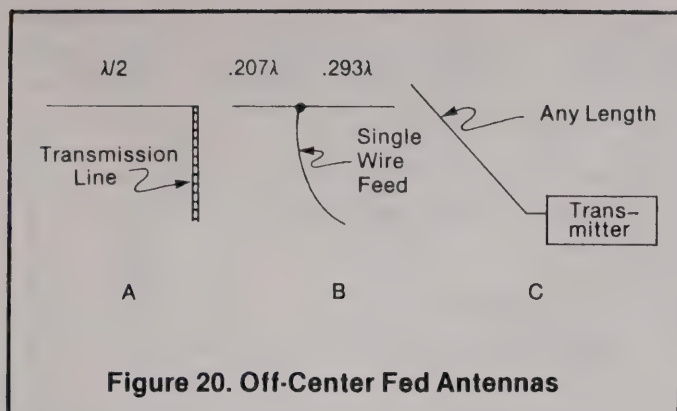
*(This is the reference dipole)

The extended double Zepp is the longest center-fed dipole with a radiation pattern remaining at right angles to the line of the antenna. (Max. energy).

(2.) Off-center Fed

Antennas and their characteristics are tabulated below:

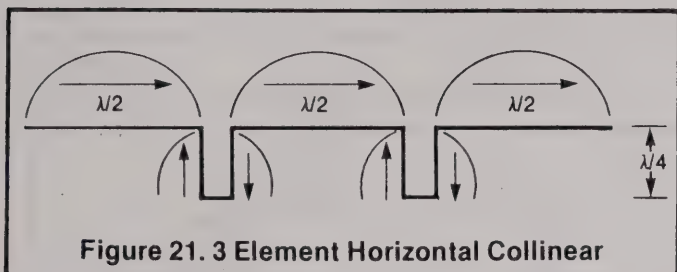
Antenna	Figure	Characteristics
Zepp	20-A	Voltage fed with balanced transmission line
Windom	20-B	Resonates on even order harmonics
Fuchs	20-C	Long wire brought directly to transmitter



b. Stubbed

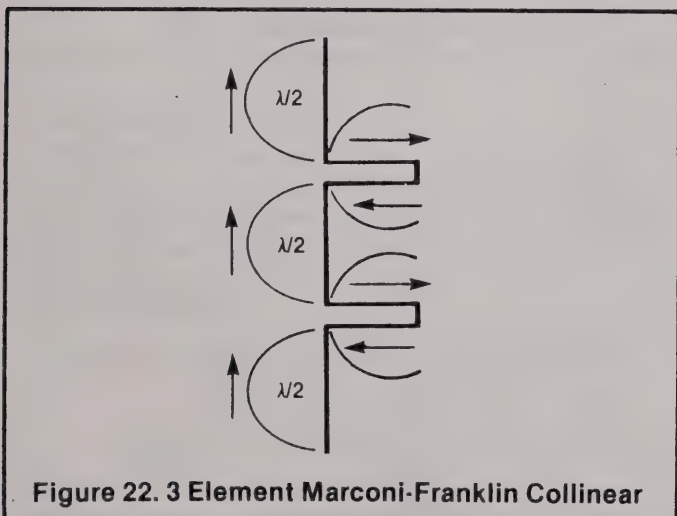
Stubbed "dipoles" are in-phase radiating elements arranged in a single line or collinear. The stub between elements insures that each successive element is fed at the same phase relationship.

A three element horizontal collinear is illustrated in Figure 21.



It is a simple antenna and can be fed at almost any point. Additional gain is achieved simply by adding another stub and half-wave element.

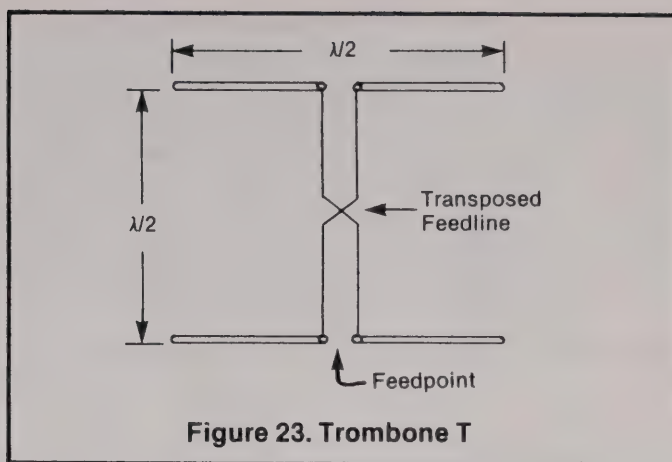
The vertical counterpart of this collinear is called a Marconi-Franklin array. (Figure 22)



In both antennas current direction is indicated by the arrows.

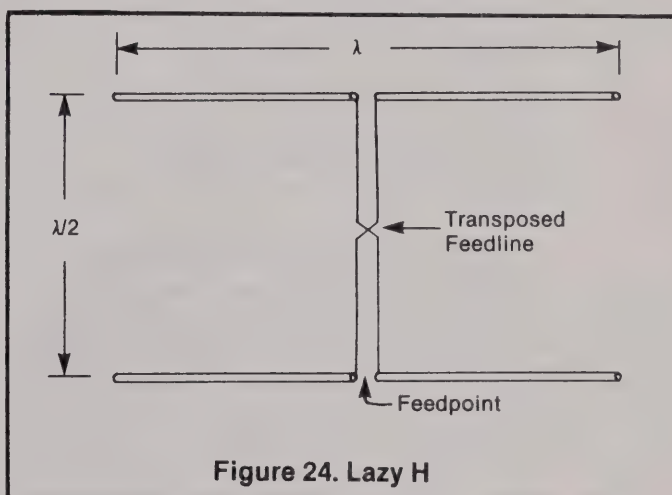
3. Trombone T

The Trombone-T is a simple 4 dBd gain broad-side antenna of low-cost. It is illustrated in Figure 23.



4. Lazy H

The Lazy-H is a twice as large version of the Trombone T. Gain is approximately 6 dBd (5.8 dBd) and it is voltage instead of current fed. See Figure 24.



● Endfire

An endfire configuration is characterized by radiation in the same plane formed by the antenna or array. It can be further subdivided into the broad categories of Frequency Dependent or Frequency Independent.

● Frequency Dependent

Antennas that operate over a narrow band i.e. maintain directional patterns and feedpoint impedance response appropriate for its intended use are considered in the classification — Frequency Dependent. They are:

1. 8 JK
2. Double Triplex Beam
3. ZL Special
4. Cardioid Beam Antenna

1. 8JK

The 8JK was named after its inventor John D. Kraus W8JK (Radio amateur call sign). It is an

endfire array with approximately 4.3 dBd bidirectional gain. It is a simple antenna to construct with reliable performance. It is illustrated in Figure 25.

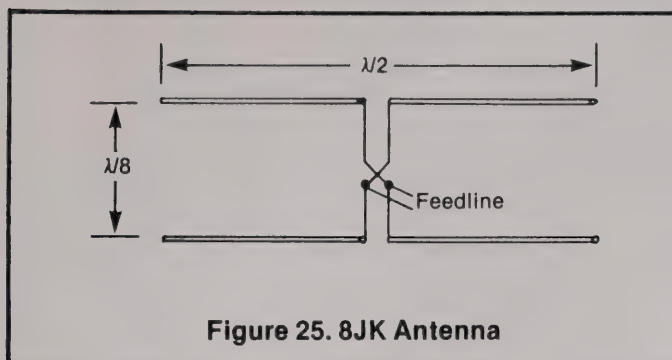


Figure 25. 8JK Antenna

2. Double Triplex Beam

The Double Triplex Beam is basically an 8JK with the single linear conductors replaced by triple leads. It has the added advantages over the single 8JK of:

- Broadband frequency response
- High radiating efficiency (300 ohm feedpoint impedance)
- Good wet weather performance. (See Figure 26.)

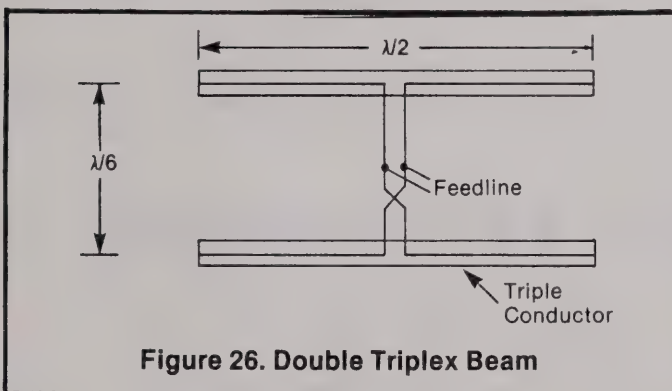


Figure 26. Double Triplex Beam

3. ZL Special

The ZL Special is a two element driven end-fire array. The elements are electrically driven 135° out of phase and are spatially phase separated by an additional 45°. It is unidirectional, has a 20 dB front to back ratio and is lightweight and easy to construct. (Figure 27)

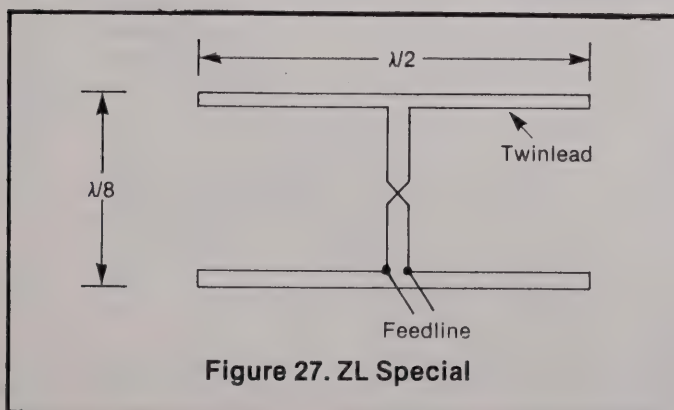


Figure 27. ZL Special

4. Cardioid Beam Antenna

A unidirectional endfire pattern is generated if the total-phase separation (electrical & spatial) between elements is 180°. (i.e. the respectively generated wavefronts must combine in 1 direction (0, 360°) and cancel in the opposite direction (180°). The Cardioid Beam Antenna develops this 180° phase separation with a 90° electrical and 90° spatial phase condition. The resulting pattern is of a broad-nosed cardioid response. (Figure 28)

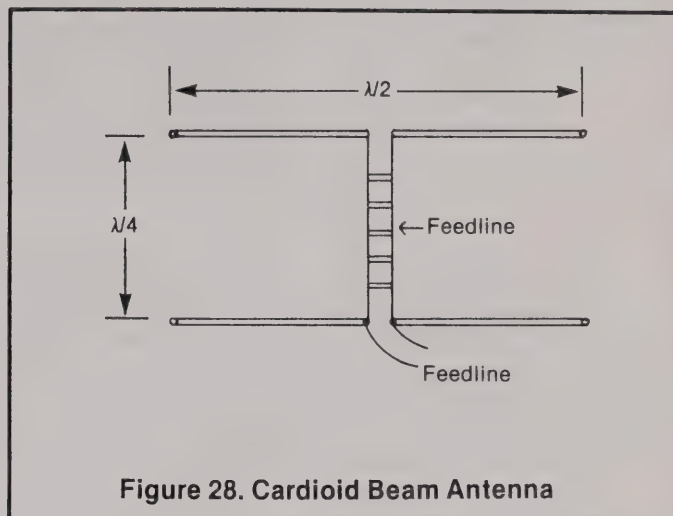


Figure 28. Cardioid Beam Antenna

● Frequency Independent

Conversely, antennas that maintain a relatively constant terminal (feedpoint) impedance and vertical angle (elevation angle) or display other frequency invariant directional properties, over several octaves, are classified as frequency independent. The logarithmically periodic array or log periodic for short is a prime example of an RF range frequency independent antenna.

Log Periodic Arrays³⁶

A Log Periodic array is a multi-element end-fire driven antenna in which adjacent elements length and spacing are equal to a constant ratio τ .

A second parameter σ is related to the ratio of the separation between adjacent elements and 4 times the length of the shorter element. An illustration and set of formulas will help clarify. See Figure 29. (On page 15).

The gain of a Log Periodic is a function of τ and σ . Other design criteria are:

$$2l_1 = 0.47 \lambda_{\max}$$

$$2l_{\min} \leq 0.38 \lambda_{\min}$$

$$R_1 - R_{\min} \geq 0.50 \lambda_{\max}$$

With these criteria and the required high and low frequencies 3 nomographs are used to complete the design.

Verticals

The simplest of vertical antennas is sometimes represented as the evolution of an open-ended

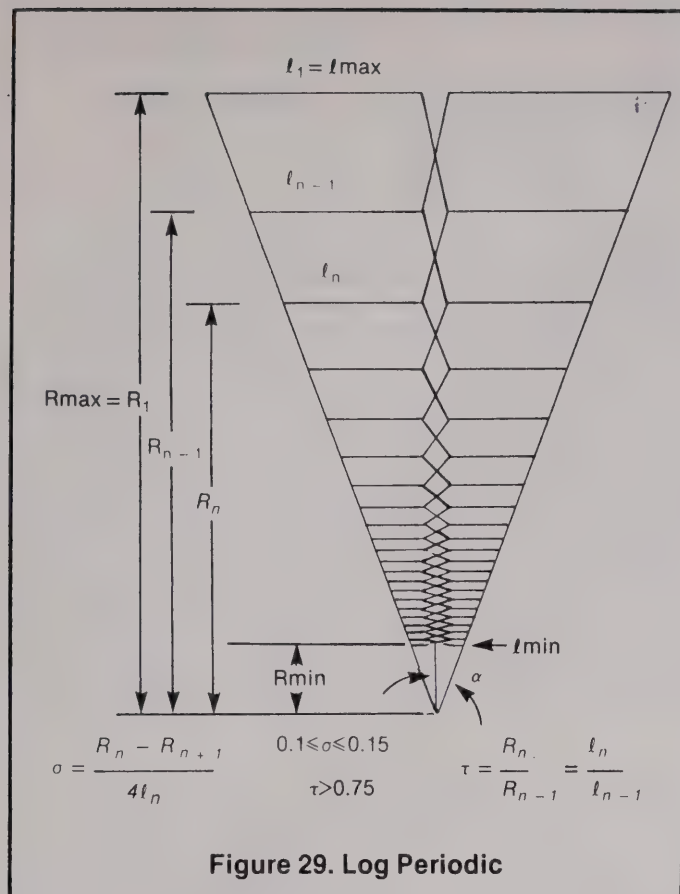


Figure 29. Log Periodic

coaxial transmission line. The outer conductor is folded back exposing the inner conductor and the combination of inner and outer conductors form a complete radiating system or antenna with specific feedpoint impedance and directional pattern.

The vertical antenna is probably the most widely recognized radiator by the public due to its prolific utilization by commercial AM broadcast stations, naval whips and hobbyists in the form of amateur and CB car and roof-mounted ground planes.

Discounting re-oriented horizontal antennas and special categories such as the stubbed collinear or Franklin array, discussed earlier, verticals can be classified as follows:

1. Single driven conductor, length $\leq 5/8 \lambda$
2. Single driven conductor, length $> 5/8 \lambda$
3. Broadbanded verticals
4. Special application verticals

1. Single Driven Conductor, Length $\leq 5/8 \lambda$

Vertical antennas are commonly utilized at the following heights: $1/4 \lambda$, $1/2 \lambda$ and $5/8 \lambda$. All three heights lay down a maximum signal strength (theoretically) at the horizon or zero elevation angle. This requires that the ground mounted vertical work against an infinite extent, infinite conductivity ground system.

The field strength, under these conditions, along the horizontal as a function of antenna height for

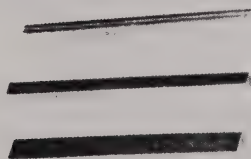
MICROWAVE ABSORBER

MATERIAL

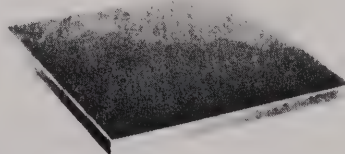
PARTS

Rigid, Flexible & Castable Materials for your -

- Coaxial & Waveguide load elements
- Attenuator elements
- Back Cavity Spurious Absorbers
- Harmonic Absorber Filters
- Rotary Joint Gaskets



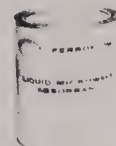
Machinable
Round & Square Bars
Waveguide ID Bars



Machinable
Plate Stock
1/8" - 1" TK & 3x3 to 12x12



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a vertical grounded radiator with one kilowatt radiated power is tabulated below:

Antenna Height	Field Strength
$\lambda/4$	196 mv/meter
$\lambda/2$	236 mv/meter
$5/8 \lambda$	276 mv/meter

Such is rarely the case. Usually, in commercial service, (AM broadcast) 120 radials are laid out every 3 degrees. This does help stabilize the impedance of the vertical system though in itself does *not* lower the maximum radiation angle to the horizon.

Simple techniques are however available for establishing an extremely low (as low as 0°) elevation or takeoff angle. Mounting a vertical above ground with a radial or counterpoise system drooping downward at about 45° helps to significantly lower the elevation angle (of maximum radiation). One precaution worthwhile mentioning is that the counterpoise should be at least a $\lambda/4$ above ground.

Numerous studies have developed working relationships between vertical antenna efficiency, length and number of radials.^{26, 27, 28, 29}

2. Single Driven Conductor, Length $> 5/8 \lambda$

Verticals greater than $5/8 \lambda$ in continuous length are difficult to construct for the lower frequencies and as a result of vertical angle lobe splitting and its concomitant high angle radiation are not as widely utilized as the previous category.

3. Broadbanded Verticals

There are several vertical antennas that operate over large bandwidths with VSWRs not exceeding 2.0:1. Two of them, the Discone and the Conical Monopole are discussed.

a. Discone

The discone is a wideband vertical radiator that can be directly matched to a 50 ohm system. The

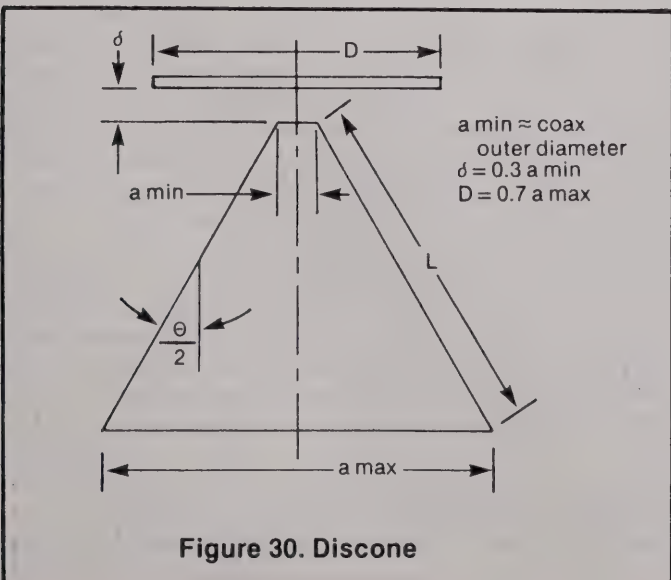


Figure 30. Discone

basic configuration of a Discone is indicated in Figure 30.

The VSWR frequency variation at the low end of the band is determined by L and θ . An optimized discone with $\theta = 60^\circ$ and $a/L = 1/22$ has a VSWR of less than 1.5:1 over a 7 to 1 frequency range and a VSWR of less than 2.0:1 over a 9 to 1 frequency range.³⁰

b. Conical Monopole

The conical monopole is short vertical radiator resembling two terminated cones placed base to base. It operates over a 4:1 bandwidth, and has high radiation efficiency. The configuration with dimensioning is indicated below.³¹ (See Figure 31)

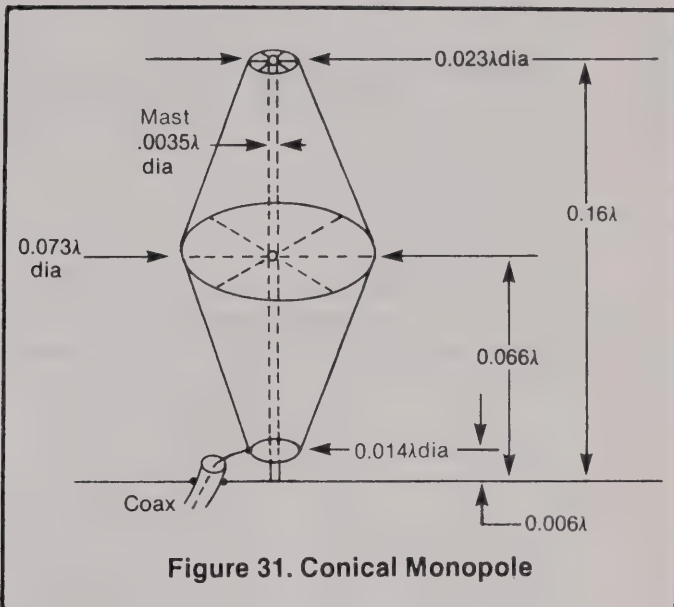


Figure 31. Conical Monopole

4. Special Application Verticals

LF and VLF applications are probably the hardest areas to accommodate in terms of antenna design due to several conflicting constraints. Foremost is the operating wavelength — in a word — long. Reasonable feedpoint impedances, bandwidth and

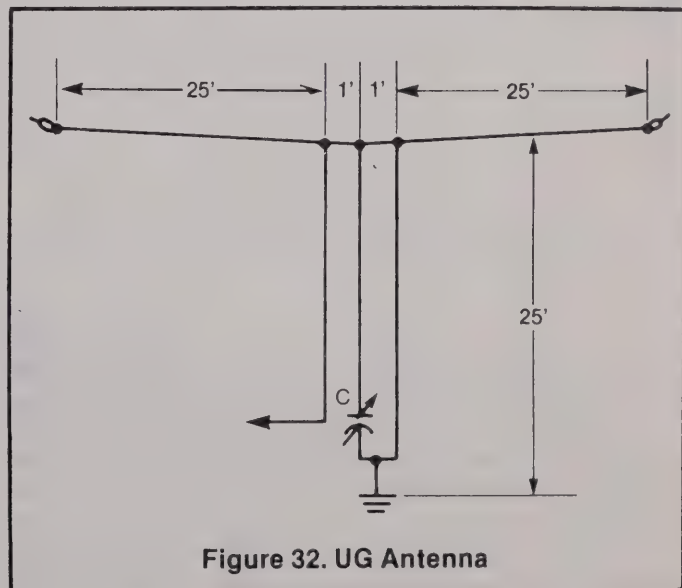


Figure 32. UG Antenna

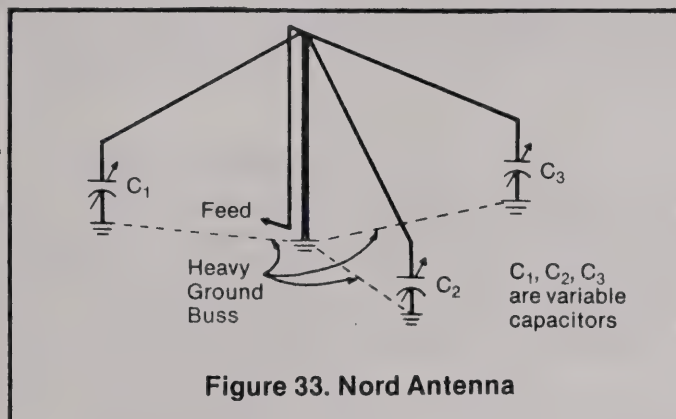
efficiencies must be maintained. Two, out of many configurations, that are in themselves tradeoffs, are discussed.

a. Type UG

The type UG antenna has the shape of a capital T as illustrated below with dimensioning given for 2 MHz operation. Capacitor C is varied until a favorable feedpoint impedance is developed.³² (Figure 32)

b. NORD

The NORD is a short vertical antenna, used at LF and MF. It has a sufficiently high bandwidth and radiation efficiency for its intended use. It is basically three over-coupled tuned circuits. A center tower is used as a common element. See Figure 33.



Leaky Wave Antennas

Leaky wave antennas can be identified by their low silhouette (close to ground) or controllable aperture distribution i.e. beamwidth and sidelobe level. Several types identified here are:

1. DDRR
2. Fast Wave
3. Slow Wave

1. DDRR

a. $\lambda/4$ DDRR

The Directional Discontinuity Ring Radiator developed by J.M. Boyer is an omni-directional low profile vertical polarization radiator. It is a single circular open-ended element mounted horizontally a short distance above a ground plane. See Figure 34.

According to theory the DDRR antenna ring forms a transmission line with its ground plane. The slot, between the ring and ground plane being a discontinuity, radiates continuously in a direction transverse to the ring axis.

The advantages of this antenna include small horizontal dimensions, extremely low height and simple construction. Since it is DC grounded induced charge static is not a problem as it is with a monopole.

b. $\lambda/2$ DDRR

A larger version of the original DDRR is available. Primary advantage is an increase in gain. It is as depicted in Figure 35.

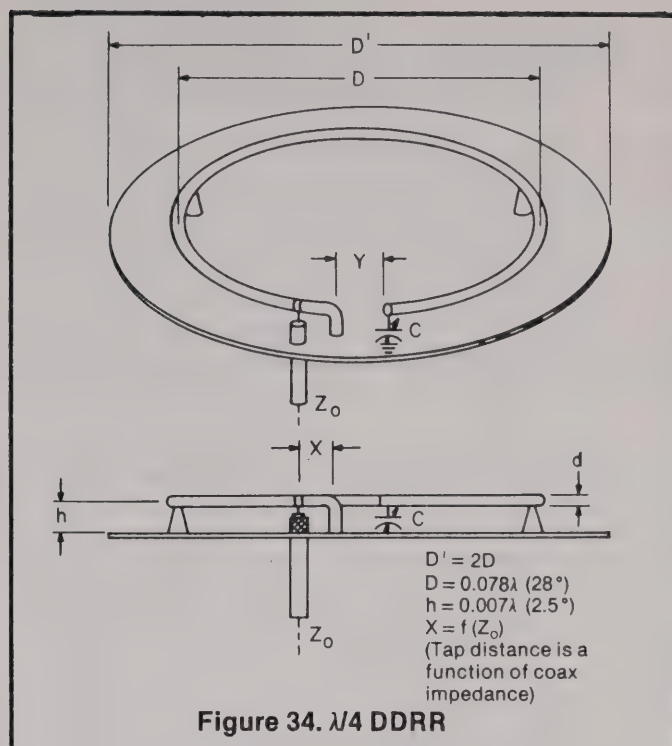


Figure 34. $\lambda/4$ DDRR

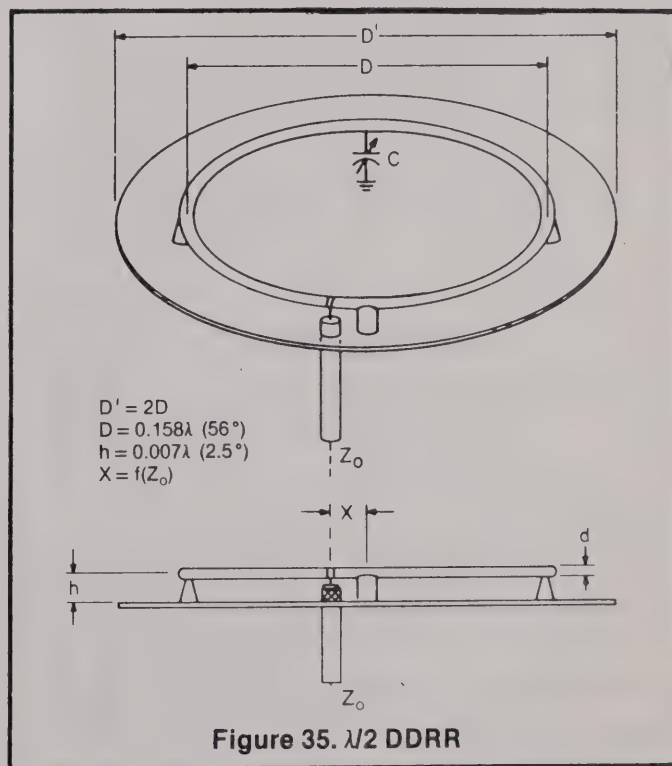


Figure 35. $\lambda/2$ DDRR

2. Fast Wave

Another class of leaky wave antenna is the Fast Wave Structure. Similar to the DDRR it is a low silhouette structure. It is cut along its length to permit the radiation of energy³⁴. The angle of radiation is closely related to the phase velocity of the guiding structure before the cut is made. The gaps must be smaller than $\lambda/3$ of the highest frequency of operation.

3. Slow Wave

An additional category of leaky wave antenna is a slow-wave structure. It is periodically-loaded (series capacitors or shunt inductors) long wire antenna. The radiation angle is determined by the phase velocity of the slow wave and the periodicity of the loading. The spacing between successive loadings must be approximately a half-wavelength or greater, for with closer spacings no radiation occurs.

Reflector Antennas

Beam formation in this type of antenna relies on the shape of the reflecting surface to focus energy from a line or point at the feed position into a narrow parallel wavefront beam. The converse also holds i.e. intercepted parallel wavefronts are focused into a common line or point location. Two types considered are:

1. Corner Reflector

A corner reflector is a half-wave radiator with a reflector consisting of two plane metal surfaces meeting at an angle immediately behind the radiator.

The radiation is the sum of direct radiation from the dipole and reflections from the planes. Consequently the pattern depends on the distance from the dipole to the corner.

At VHF the corner reflector is often solid (metal). At lower frequencies screen wire or metal rod constructed reflectors are used. The maximum gap between the wires or rods should not exceed 0.06λ .

Power gains of at least 10 dB to as much as 14 dB are achievable using a corner reflector. A diagram with design information is shown in Figure 36.

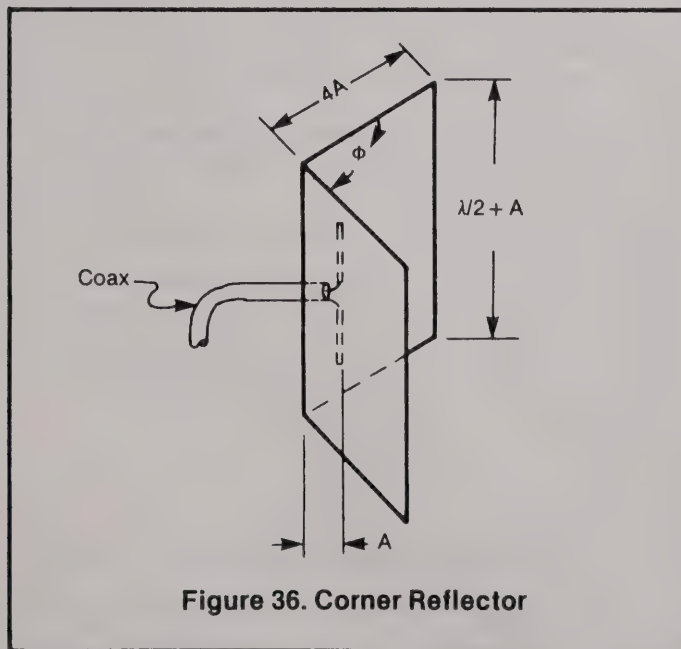


Figure 36. Corner Reflector

Principal advantages of the corner reflector are its simple construction and the low side and back lobes of the directivity pattern.

2. Parabolic Reflectors

Parabolic reflectors are widely used in UHF and microwave radar and point-to-point communications systems to achieve antenna gains in excess of 20 dB.

Recently the CATV industry has been successfully utilizing a Half-bolic to receive very weak over the horizon VHF signals.³⁵ (A Half-bolic, as its name implies is physically half a parabolic reflector. When combined with its ground image its performance is similar to a parabolic reflector.)

In the low frequency application an offset feed is employed to reduce undesired interaction between reflector and feed and to minimize aperture blocking.

A suitable feed might consist of a log periodic array in which the longest element forms the dipole feed of a corner reflector. The phase center of the corner reflector would be located at the focus of the parabola. Using this technique the maximum possible gain and narrowest possible beamwidth are achieved at the lowest operating frequency.

Though 5 types of antennas with 51 different configurations were discussed this is not to imply that these are, or were the only antennas ever designed. This is a broad sample of most common antenna configurations utilized by all services — commercial, government and hobbyist. Modifications such as shortening, employing active elements and material experimentation affects the properties and probably the classifications of the antennas already discussed.

Future articles will delve into specialized antenna configurations utilizing some of the above-mentioned techniques, and others not yet considered. □

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This nested rhombic is fed by 300 ohm balanced line, is mounted 12.29 feet above ground and has a vertical and horizontal beamwidth of 5.5° and 8.5°, respectively.

● Close to the Ground ($\leq \lambda/8$)

● Beverages

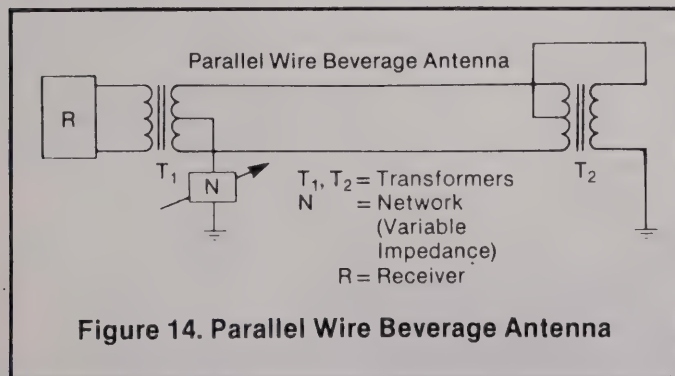
A Beverage antenna is a horizontal longwire antenna utilized especially for reception of low-frequency vertical polarized waves. It normally consists of a single wire several wavelengths long erected close to the ground, or even on the ground, and terminated at the far end in its characteristic impedance.

An outstanding merit of this antenna is its ability to "pull signals out of the noise". The low end of the HF spectrum is typically atmosphere noise limited as opposed to thermal noise limiting in the higher RF ranges. Most HF antennas previously described present large signal *and* noise voltages at receiver terminals. The Beverage, on the other hand, generates a lower signal voltage while at the same time a considerably lower noise voltage. The overall effect is to produce a better or higher signal-to-noise ratio which is really the name of the game in communications intelligibility.

● Parallel Wire Beverage Antenna

An interesting and very useful feature related to the operation of a Beverage is the ability to reject signals not only off the back but off the sides — selectively. By tuning the complex termination impedance depth and direction of rejection can be adjusted.

A twin or parallel wire Beverage places the termination network close to the receiver (and operator). (Figure 14)



T₁, T₂ = Transformers
N = Network (Variable Impedance)
R = Receiver

Several comprehensive research efforts have explored many aspects of Beverage antenna design,

development, deployment and measurement.^{20, 21, 22, 23, 24} They are well worth reading about.
Part 2 in the series continues with "Driven Antennas."

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Trends in HF Receiving Antennas

CYRIL WHITBREAD
C & S Antennas Ltd.
Rochester, Kent, UK

Only a few years ago many people thought that HF communications was in a decline. The proliferation of communication satellites and the greatly increased number of circuits made available by submarine cables suggested that HF radio would be quickly abandoned. Today, however, HF still holds a strong position and manufacturers throughout the world are turning their expertise to improving HF performance at all points of the link.

There are several reasons for this return to popularity. For developing countries, HF is a quick and cheap way into international communications. The ground wave propagation of HF also makes it effective in tactical use, and there is a growing realization of the important role which

HF can play in broadcasting. The overriding fact is that the demand for communications is growing more rapidly than any of the more recent techniques can satisfy; while that situation remains, HF will continue to flourish.

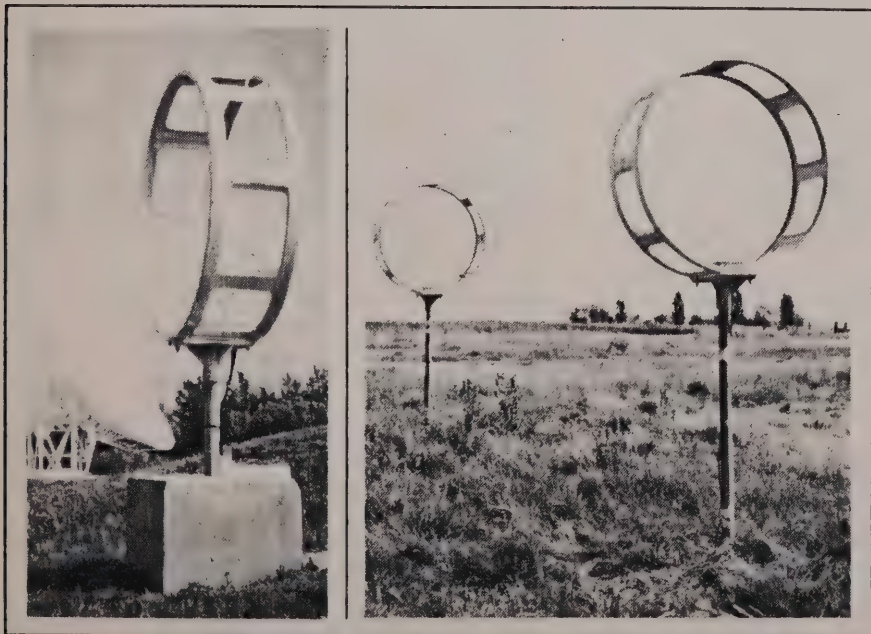
Despite this popularity, HF is not without its problems. At the receiving end, one of the problems is noise. With VHF and UHF noise is generated by the electronic equipment in the system, but with HF the noise comes in through the antenna from thunderstorms, galactic disturbances, and other external sources. Increasing the antenna gain does not provide a solution, because this only increases noise as well. What is needed is discrimination against the noise.

Discrimination can be achieved if, instead of having an antenna which looks at all angles, one has an antenna which looks only at the point from which the signal is emanating. In this way, reception of the majority of the noise can be eliminated. Directivity is, therefore, the significant factor, not signal amplitude. The real need for successful HF reception is a highly directive antenna.

The rhombic antenna is effective for long-distance HF point-to-point communications, but while it has very good directivity, it usually produces more signal than the receiver needs and often more than it can cope with. It is also a fairly complex structure which calls for time-consuming fabrication and assembly on site. The result is a large, permanent installation covering a good deal of ground.

None of these drawbacks applies to the active loop antenna, which, during the past few years, has been employed to a growing extent in a variety of HF receiving installations. Most people think of loops as being bi-directional, giving a figure-of-eight radiation pattern. However, bearing in mind that directivity is the important thing, it has been possible to design a loop antenna which is uni-directional.

The directivity of a single loop is not very high, so the user is going to have to work hard to gain much advantage over a simple wire antenna. But when the loops are built into arrays, directivity is enhanced and the signal-to-noise ratio is increased. The arrays can be steered by switching, so it becomes possible to manipulate the directivity.



A typical array of active loop antennas.

A typical, practical realization of the active loop antenna takes the form of a 92 cm-diameter broad loop of flat aluminium alloy encased in glass fibre, mounted on a pole (Figure 1). This structure weighs only 19 kg, and the centre of the loop is typically only 1.5 m above the ground. A transistorized amplifier is mounted in a recess at the bottom of the loop, and in operation the antenna follows the classical concept of the directional coupler. The loop is joined at the top by a resistor of exactly the right value, which allows the current due to a field travelling in one direction to add together, and the current due to a field travelling in the other direction to cancel out, so giving directivity.

Each loop is vertically polarized, with a frequency range of 2-30 MHz, directivity of 3 dB, and front-to-back ratio of 12 dB. The amplifier has been designed to have noise and intermodulation performance which will permit it to operate over the entire HF band under low signal level conditions.

The active loop antenna can be used singly, but the real benefits are gained when the loops are placed in arrays. As few as four loops can make an effective array, although much better performance is achieved if an array of eight loops is deployed. With eight or more loops, the variety of arrays which can be constructed is very wide.

End-fire and broadside arrays are both possible. With the end-fire array, a single azimuth directional beam is provided with low sidelobe levels and a high back-to-front ratio maintained over a wide frequency band. Correct spacing of the loops optimizes the performance of the array for the required frequency band. Choice of beam elevation angle is possible, from very low angles for long-distance circuits, to near vertical incidence for short circuits. This elevation angle may be fixed or can be adjusted by switching.

Broadside arrays are characterized by a beam which is wide in the elevation plane and narrow in the azimuth plane. The vertical radiation pattern is not dependent on the number of elements, and its beamwidth changes little with frequency. The horizontal radiation pattern, on the other hand, varies widely with frequency, the beamwidth being inversely proportional to frequency. Suitable array design allows the beam maximum to be slewed up to 45° from the normal. A single slewed beam, switched slew angles, or a

number of simultaneous outputs to different receivers can all be catered for in the design.

Installations can be designed to combine the properties of both end-fire and broadside arrays by arranging the elements in two dimensions, ranks, and files.

High directivity is not, of course, what the communications engineer is always looking for. He or she might want to scan a wide area and then concentrate on one direction. This requirement can be met by arranging the active loop antennas in a circular array, with each loop firing radially outwards. The connections are commutated round the ring, so that the beam may be switched progressively in bearing. As an alternative, multiple beam arrays can be designed to give the facility of uni-directional beams operating simultaneously in any chosen direction.

Yet another example of the versatility of active loop antennas is the log periodic array, in which the loops are positioned with a spacing that varies in accordance with a geometric progression. In this case, the array combines the advantages of small size and ease of installation offered by wide-band active antennas with those of log periodic technology - high directivity and beamwidth which is constant with frequency.

Even greater directivity can be achieved by arranging two log rows of loops in a V-shaped array. The antenna designer need not stop there, for additional log rows of elements can be arranged in a rosette pattern to make a succession of V formations around a central point, so that all bearings are covered.

Quite apart from its good performance characteristics and versatility, the active loop antenna offers a number of practical advantages. The amount of space taken by an installation can be small; it is possible, for example, to place an array of four loops on the roof of a building. The low profile of the loops makes them acceptable environmentally, and as each element can be handled easily, the installation of an array is much easier than with a traditional antenna structure.

A simple installation on a roof or in a field can be completed in an afternoon by a few unskilled personnel, and even the largest and most complex arrays involve none of the terrors of traditional wire antenna rigging.

Normally the pole of the loop antenna is set into a concrete block, but

the relatively small size and low weight of the loops means that they readily lend themselves to temporary installations, and various alternative mountings can be supplied for this purpose. Any flexible 50 ohm coaxial feeder may be used to connect each loop in an array to the combining network. For military use, a transportable version of the loop antenna is available which folds like a deck chair and can be deployed in a matter of seconds.

The savings in "real estate" offered by the active loop antenna are not limited to single arrays. As mutual coupling between loops is very small, it is possible to interleave arrays on the same site. For example, a site measuring only 50 by 40 m could be occupied by five arrays, receiving signals from different bearings and operating at different frequencies. This would be extremely difficult to achieve by conventional means, but severe crowding presents few difficulties with active loops. Not only can the antennas be placed very closely together, multiple beam methods can also be used to reduce the number of arrays needed.

Maintenance is also simple. No climbing is necessary, as the amplifiers are mounted only about one metre above the ground. The performance of each amplifier can be monitored, and there is normally a high degree of redundancy in an array, so that even if one amplifier fails, the array remains operational. Practical experience now confirms that the low profile of loop arrays makes them virtually impervious to lightning strikes, allowing them to continue in operation when neighbouring rhombics have suffered substantial damage.

For the future, there can be no question that HF communications will continue to be widely used, and even grow in popularity. The active loop antenna can answer most problems of reception and its advantages of performance, versatility, limited site requirements, and simplicity should ensure that it is used to an increasing extent in HF receiving systems.

Military users can clearly benefit, and the portability and ease of installation of the active loop also make it ideal for emergencies, such as natural disasters, or other occasions when communications are needed urgently. But the range of applications is very wide, and all those requiring high-grade HF receiving systems are likely to look on the active loop antenna with more and more favor. (ES)

Introduction to radio and radio antennas

The concepts of wave transmission can be more readily grasped if the approach is based upon an understanding of the fundamental circuit elements: inductors and capacitors

Harald T. Friis Rumson, N.J.

Electromagnetic wave propagation is a simple and beautiful phenomenon, but its rigorous mathematical derivation from Maxwell's equations is formidable. This article makes many of the concepts visible and plausible using only mathematics available to college freshmen and taught today in most high schools. Those who are familiar with the subject will appreciate the simple and direct way in which Dr. Friis' transmission formula is derived. This derivation is new and is published here for the first time.

—David DeWitt, Editor

Transmission lines guide telephone signals, and also power, from one place to another. Although such guidance does not exist for radio signals, it is illustrative to use the transmission of waves between two parallel conductors as an introduction to radio.

A transmission line has a capacitance C and an inductance L , per unit length, and in order to derive the transmission-line equations we must know the fundamental properties of the important circuit elements: inductors and capacitors.

Inductors

Figure 1 shows a long coil of wire with direct current i . The resistance of the wire is assumed to be negligible. By means of a magnetic needle indicator it has been found that the magnetic field H inside the coil is constant. Physicists also found for a closed path around currents that the sum of the magnetic fields along the path multiplied by the path lengths is equal to the total current through the area enclosed by the path. Path $ABCD$ in Fig. 1 shows such a path. The magnetic field is zero along this path except along AD , and the total current through the area of the path is $i \times n$. Thus,

$$Hl = in$$

Editor's note: Last month an autobiographical memoir by Dr. Friis, *Seventy-Five Years in an Exciting World*, was published by San Francisco Press, 555 Howard St., San Francisco, Calif.; it also contains "The Wisdom of Harald T. Friis," based on a famous speech (annotated by John R. Pierce) in which the author distills a lifetime's hard-won research experience.

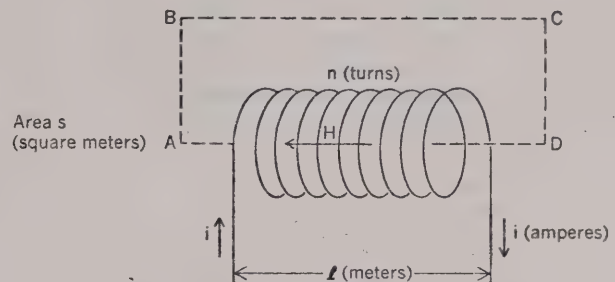
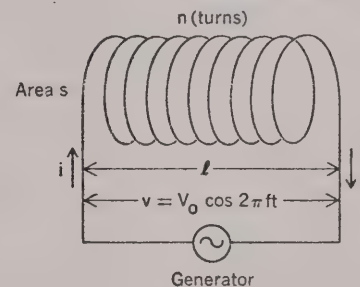


FIGURE 1. Long coil of wire with direct-current flow.

FIGURE 2. Long coil with ac applied voltage.



or, by definition,

$$H = \frac{in}{l} \text{ amperes per meter}$$

In Fig. 2 an ac voltage $v = V_0 \cos 2\pi ft$ is applied to the coil; f is the frequency in hertz and t is the time in seconds. (See Fig. 3 for variation of v vs. $2\pi ft$.) Faraday found that the voltage induced in one turn by i is proportional to the area s of the turn and the rate of change of i or H . Hence,

$$v \text{ (one turn)} = \mu s \frac{dH}{dt} = \mu s \frac{n}{l} \frac{di}{dt}$$

where the proportionality factor μ is the permeability of the medium inside the coil (for air, $\mu = 4\pi \times 10^{-7}$ H/m). The voltages in the separate turns are aiding. Therefore, the voltage induced across the coil is

$$v = \mu s \frac{n^2 di}{l dt} = L \frac{di}{dt} \text{ volts} \quad (1)$$

$L = \mu s(n^2/l)$ is called the inductance of the coil and is measured in henrys.

For $v = V_0 \cos 2\pi ft$, Eq. (1) may be solved by integration to obtain

$$i = \frac{V_0}{2\pi fL} \sin 2\pi ft \text{ amperes} \quad (2)$$

The values of i and v are plotted versus $2\pi ft$ in Fig. 3.

Capacitors

Figure 4 shows the plates of a capacitor. Electrostatic experiments showed that the ratio of total charge q on each plate to the voltage v between the plates is constant:

$$\frac{q}{v} = \text{constant} = C \quad (3)$$

C is proportional to the ratio of the area s (square

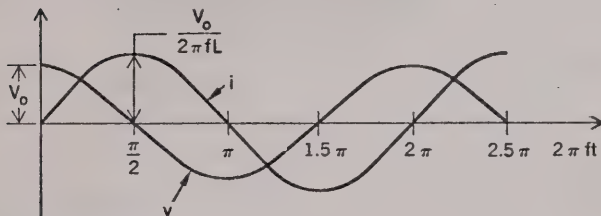


FIGURE 3. Voltage (v) and current (i) vs. $2\pi ft$ for inductive circuit of Fig. 2.

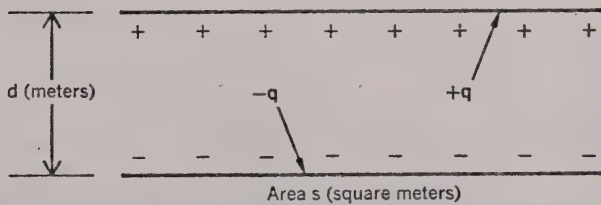


FIGURE 4. Basic capacitor configuration.

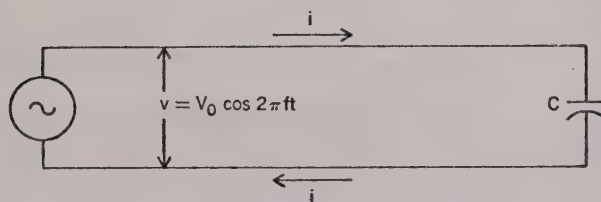
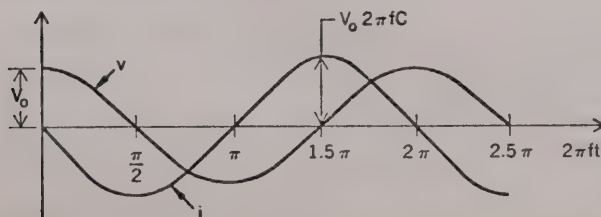


FIGURE 5. Capacitor with ac applied voltage.

FIGURE 6. Voltage (v) and current (i) vs. $2\pi ft$ for capacitive circuit of Fig. 5.



meters) and spacing d (meters) of the plates:

$$C = \epsilon \frac{s}{d} \text{ farads} \quad (3a)$$

where ϵ , the proportionality factor, is the dielectric constant of the medium between the plates. For free space $\epsilon = (1/36\pi)10^{-9}$ F/m.

In Fig. 5 a voltage $v = V_0 \cos 2\pi ft$ is applied to the plates. The current i , called the displacement current, is equal to the rate of change of q . Equation (3) gives

$$i = \frac{dq}{dt} = C \frac{dv}{dt} = -2\pi fCV_0 \sin 2\pi ft \quad (4)$$

The values of i and v are plotted versus $2\pi ft$ in Fig. 6.

Transmission lines

Figure 7(A) shows a generator that feeds a long, loss-free uniform line with an inductance of L henries and capacitance of C farads per meter and Fig. 7(B) shows the equivalent lumped-circuit line.

The current i_x through inductor $L dx$ decreases the voltage from AB to CD . Equation (1) gives

$$dv_x = -L dx \frac{di_x}{dt} \quad (5)$$

or

$$\frac{dv_x}{dx} = -L \frac{di_x}{dt}$$

Differentiating with respect to t gives

$$\frac{d^2v_x}{dx^2} = -L \frac{d^2i_x}{dt^2} \quad (5a)$$

The displacement current through capacitor $C dx$ decreases the current from A to C . Equation (4) gives

$$di_x = -C dx \frac{dv_x}{dt} \quad (6)$$

or

$$\frac{di_x}{dx} = -C \frac{dv_x}{dt}$$

Differentiating with respect to x gives

$$\frac{d^2i_x}{dx^2} = -C \frac{d^2v_x}{dx dt} \quad (6a)$$

Substituting the value for $d^2v_x/dx dt$ given by Eq. (5a) gives

$$\frac{d^2i_x}{dx^2} = LC \frac{d^2i_x}{dt^2} \quad (7)$$

Similarly, we obtain

$$\frac{d^2v_x}{dx^2} = LC \frac{d^2v_x}{dt^2} \quad (8)$$

Mathematicians can solve for i_x and v_x in Eqs. (7) and (8) and get, as one correct solution,

$$i_x = I_0 \sin(2\pi ft - 2\pi fx\sqrt{LC}) \quad (9)$$

$$v_x = V_0 \sin(2\pi ft - 2\pi fx\sqrt{LC}) \quad (10)$$

We can verify that these solutions are correct by differentiating i_x in Eq. (9) twice with respect to x and twice with respect to t :

$$\frac{di_z}{dx} = -I_0 \cos(2\pi ft - 2\pi fx\sqrt{LC}) \times 2\pi f\sqrt{LC}$$

and

$$\frac{d^2 i_z}{dx^2} = -I_0 \sin(2\pi ft - 2\pi fx\sqrt{LC}) \times (2\pi f\sqrt{LC})^2 \quad (11)$$

Also,

$$\frac{di_z}{dt} = I_0 \cos(2\pi ft - 2\pi fx\sqrt{LC}) \times 2\pi f$$

and

$$\frac{d^2 i_z}{dt^2} = -I_0 \sin(2\pi ft - 2\pi fx\sqrt{LC}) \times (2\pi f)^2 \quad (12)$$

Equations (11) and (12) give

$$\frac{d^2 i_z}{dx^2} = LC \frac{d^2 i_z}{dt^2} \quad (13)$$

Equations (7) and (13) are identical. Therefore, Eq. (9) gives a correct solution of (7) and, similarly, (10) gives a correct solution of (8).

We substitute from Eqs. (9) and (10) in (5):

$$\begin{aligned} -2\pi f\sqrt{LC}V_0 \cos(2\pi ft - 2\pi fx\sqrt{LC}) \\ = -LI_0 2\pi f \cos(2\pi ft - 2\pi fx\sqrt{LC}) \end{aligned}$$

That is,

$$\sqrt{LC}V_0 = LI_0$$

or

$$V_0 = I_0 \sqrt{\frac{L}{C}}$$

From Eqs. (9) and (10),

$$v_x = i \sqrt{\frac{L}{C}}$$

or the impedance of the line is

$$Z = \frac{v_x}{i_x} = \frac{V_0}{I_0} = \sqrt{\frac{L}{C}} \text{ ohms} \quad (13a)$$

Values of v_x , as given by Eq. (10), versus x are plotted in Fig. 8 for $t = 0, t = 1/4f, t = 1/2f, \dots$

Note that the first voltage maximum moves with time from A to B to C to D to E to F with a speed

$$c = \frac{1}{2f\sqrt{LC}} \bigg/ \frac{1}{2f} = \frac{1}{\sqrt{LC}} \text{ m/s}$$

or the input generator causes a traveling wave on the line.

The separation between voltage maximums is called the wavelength λ :

$$\lambda = \frac{1}{f\sqrt{LC}} = \frac{c}{f} \text{ meters}$$

Equation (10) shows that the difference in phase ϕ of the voltage at distance x and distance $x + d$ is

$$\begin{aligned} \phi &= 2\pi f(x + d)\sqrt{LC} - 2\pi fx\sqrt{LC} \\ &= 2\pi fd\sqrt{LC} = 2\pi \frac{d}{\lambda} \end{aligned} \quad (14)$$

Figure 9 shows a uniform strip transmission line in

which the edge effect is neglected. The transmission-line theory gave line impedance $z = \sqrt{L/C}$ ohms.

In this line the inductance L per meter is the inductance of a single-turn coil with area $s = 1h$ and length $l = a$. Equation (1) gives

$$L = \mu h \frac{1}{a}$$

Similarly, Eq. (3a) gives

$$C = \epsilon \frac{a}{h}$$

The speed of propagation c is given by

$$c = \frac{1}{\sqrt{LC}} = \frac{1}{\sqrt{\mu\epsilon}}$$

In free space,

$$\mu = 4\pi \times 10^{-7} \quad \text{and} \quad \epsilon = \frac{1}{36\pi} \times 10^{-9}$$

so $c = 3 \times 10^8$ m/s, which is the speed of light and also the propagation speed for transmission lines constructed

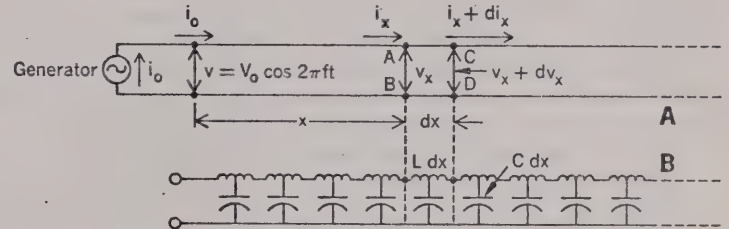


FIGURE 7. A—Long uniform transmission line, B—Equivalent lumped-circuit line.

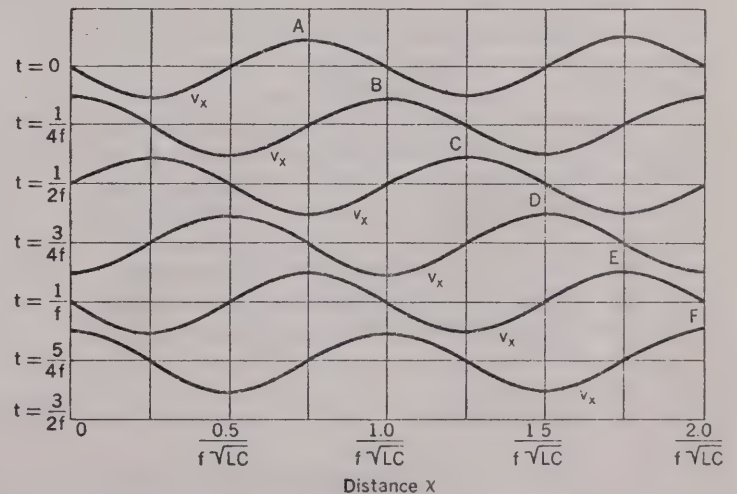
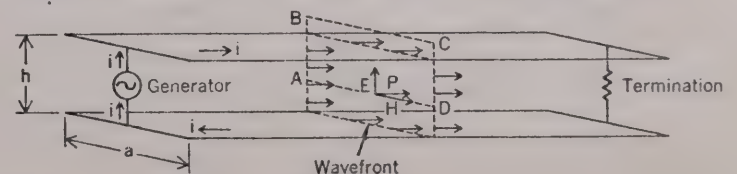


FIGURE 8. Line voltage v_x vs. distance x along the line.

FIGURE 9. Uniform strip transmission line with traveling wave.



with negligible amounts of insulation and magnetic material.

The line impedance z is given by

$$z = \frac{h}{a} \sqrt{\frac{\mu}{\epsilon}} = 120\pi \frac{h}{a}$$

For $h = a = 1$ and neglecting edge effects, we get Schelkunoff's formula for the impedance of free space:

$$z_{\text{free space}} = 120\pi \text{ ohms}$$

A sheet with 120π ohms between two opposite edges of a square of the sheet is called a resistance sheet with 120π ohms per square. It matches waves in free space and, if inserted between the ends of the line, makes a good termination; that is, it causes no reflected waves for the traveling wave on the line when $h < \lambda$. (A conducting sheet $\lambda/4$ behind the resistance sheet is required when h is larger.)

The voltage across the line is hE volts. As shown in Fig. 9, E is the maximum value of the electric field in the space between the strips. It is measured in volts per meter. The power absorbed by a circuit of impedance Z with an input voltage E is $\frac{1}{2}E^2/Z$. The impedance of this transmission line is $Z = 120\pi h/a$. The power flow in the line is, therefore,

$$\frac{\frac{1}{2}(hE)^2}{z} = \frac{\frac{1}{2}E^2}{120\pi} ha \text{ watts}$$

The power flow per unit area or the power intensity in the wave between the strip is then

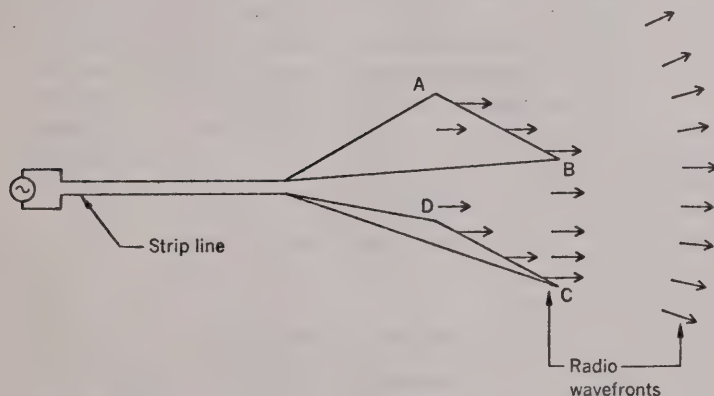
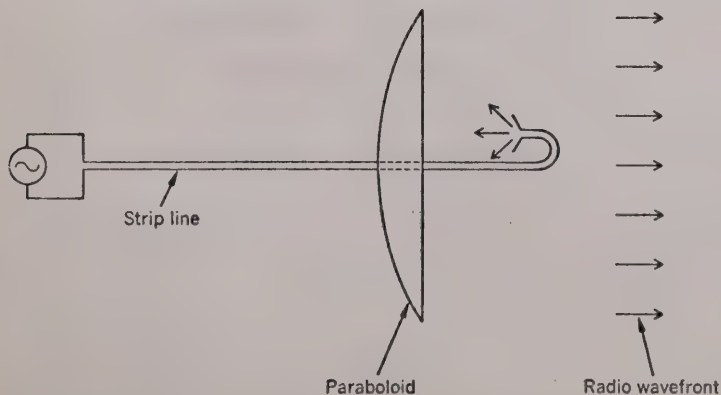


FIGURE 10. Strip line expanded.

FIGURE 11. Strip line feeding a paraboloid.



$$\frac{1}{2} \frac{E^2}{120\pi} \text{ watts (same for radio waves)} \quad (14a)$$

Note that all the power flows in the air between the strip conductors. This is true for all transmission lines. The conductors act only as guides for the waves.

Figure 9 also shows the magnetic field vector H . Its magnitude is found by choosing a path $ABCD$ and adding the field times path length as explained in the first section on inductors. The current through the areas of this path is i and the field is zero except along path AD . Hence,

$$Ha = i$$

Equation (13a) gives

$$i = \frac{v}{\sqrt{L/C}} = \frac{Eh}{120\pi h/a} = \frac{Ea}{120\pi}$$

Therefore,

$$H = \frac{i}{a} = \frac{E}{120\pi}$$

Figure 9 shows the power intensity vector P , also called the Poynting vector, from (14a),

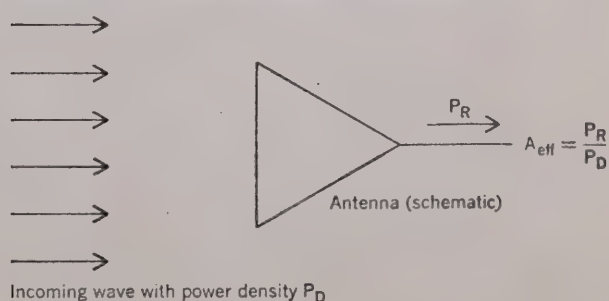
$$P = \frac{\frac{1}{2}E^2}{120\pi} = \frac{1}{2}EH \text{ watts per square meter}$$

Recapitulating what happens in the idealized line illustrated in Fig. 9, it has been shown (1) that the available power from the generator has been changed to a plane electromagnetic wave that moves with the speed of light toward the termination of the line, (2) that the power flow intensity is constant over the cross section of the wave and equal to $\frac{1}{2}EH$ watt where E is the electric field and H the magnetic field of the wave, and (3) that the impedance of the line is $120\pi h/a$ ohms and the free-space medium between the strips is 120π ohms.

The strip-line dimensions of width a and separation h may be increased gradually as shown in Fig. 10. The traveling wave will continue through this expanded section and produce a large radiating wavefront at the opening $ABCD$. We have now a radio transmitting antenna (Schelkunoff's paddle antenna). Alternatively, the strip line may feed a paraboloid, as shown in Fig. 11. We have, in other words, illustrated the passing of guided waves to radio waves; we shall next discuss some important properties of radio antennas and radio transmission.

Effective area of an antenna. The effective area A_{eff} of an antenna is defined as shown in Fig. 12, for the case of receiving, as the available power P_R from the antenna

FIGURE 12. The effective area of an antenna.



divided by the power density P_D of the incoming wave:

$$A_{\text{eff}} = \frac{P_R}{P_D}$$

If the illumination of the antenna when used for transmitting is uniform (that is, if the electromagnetic field is constant and in phase over the plane of the aperture and zero outside), the effective area is by reciprocity equal to the geometrical area of the aperture.

Examples: The effective area of the paddle antenna shown in Fig. 10 is approximately equal to the opening area $ABCD$ and the effective area of the paraboloid antenna shown in Fig. 11 is about two thirds of the opening area of the paraboloid. The effective area of a small loss-free dipole, which will be calculated later, is $(\lambda/2) \times (\lambda/4)$. As usual, all linear dimensions are in meters.

Propagation loss between two antennas in free space. That light propagates as a wave was suggested some 300 years ago by Huygens, and resulted in Huygens' principle: "Every part of a wavefront can be regarded as a source of disturbance that emits a spherical wavelet." Fresnel made his great contribution some 150 years later by realizing that the relative phase ϕ between two wavelets can be calculated from their path difference d and wavelength λ [compare with Eq. (14)]:

$$\phi = 2\pi \frac{d}{\lambda}$$

Figure 13 shows how Fresnel graphically found the field due to the wavelets in a plane wave at a distant point. Figure 13(B), a side view of the wave, indicates the wavelets on the wavefront that radiate spherically toward the distant point P . Figure 13(A) shows how the wavefront is divided into equal-area rings 1, 2, 3, $\dots$. The Huygens sources in each ring have the same distance to point P and are therefore in phase at P . Each of the rings has the same area or the same number of wavelets and therefore produces the same fields $E_1, E_2, E_3, \dots$ at P when the angle ϕ is small.

Figure 13(C) shows how two parallel fields

$$E_1 = E \cos(2\pi ft - \phi_1)$$

and

$$E_2 = E \cos(2\pi ft - \phi_2)$$

can be added graphically by plotting $OA = E$ first and then $AB = E$ so that the angle between AB and the direction of OA is $\phi_2 - \phi_1$. The sum of E_1 and E_2 is then OB . That this is correct can be shown as follows:

$$\begin{aligned} E_1 + E_2 &= E \cos(2\pi ft - \phi_1) + E \cos(2\pi ft - \phi_2) \\ &= E[\cos(2\pi ft - \phi_1) + \cos(2\pi ft - \phi_2)] \end{aligned}$$

The formula for the addition of two cosine functions gives

$$E_1 + E_2 = 2E \cos\left(2\pi ft - \frac{\phi_1 + \phi_2}{2}\right) \cos \frac{\phi_2 - \phi_1}{2}$$

or the amplitude of $E_1 + E_2$ is

$$2E \cos \frac{\phi_2 - \phi_1}{2}$$

and this is the length of OB in Fig. 13(C).

The fields at point P from the wavelets can now be added graphically, as shown in Fig. 13(D). Starting at point O , the field E_1 is plotted first with arbitrary length and direction and it is followed by E_2 so that E_1 and E_2 form an angle

$$\Delta\phi = 2\pi \frac{d_2 - d}{\lambda}$$

where d_2 is the distance from the sources in ring 2 and d the distance from the sources in the center element 1. The fields $E_3, E_4, \dots$ are plotted by continuing in the same manner. The small phase angles $\Delta\phi$ are alike when angle ϕ is small and the ends of the fields therefore follow a circle. This circle turns into a spiral when angle ϕ increases and finally gives the original field E of the wave when the rings expand to infinity.

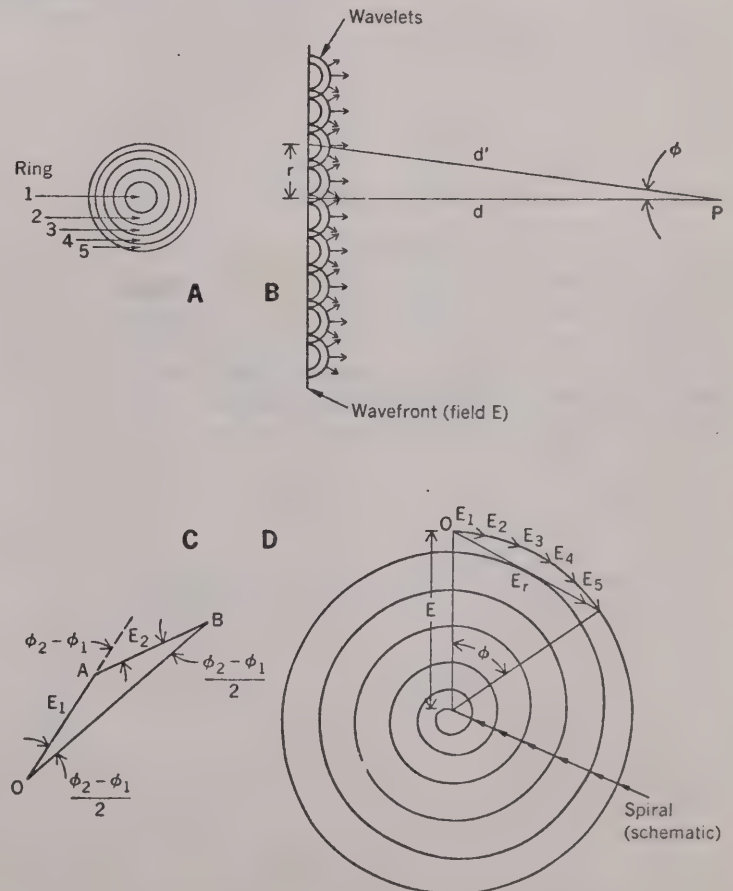
The phase of the field from the ring with radius r with respect to field E_1 from the field from the center element is

$$\phi = \frac{2\pi(d' - d)}{\lambda} \approx \frac{\pi r^2}{\lambda d}$$

Fresnel could have found the field $E_r = E_1 + E_2 + E_3 + E_4 + E_5$ due to all the rings with radii less than r . Figure 13(D) shows that it is

$$E_r = E\phi = E \frac{\pi r^2}{\lambda d} \text{ volts}$$

FIGURE 13. Propagation of light waves. A—Front view of wave. B—Side view of wave. C—Addition of two fields. D—Addition of wavelet fields.



or

$$\frac{E_r}{E} = \frac{\pi r^2}{\lambda d} *$$

This important formula will now be used to find the ratio of the power received by a light absorber with

* Since E_r (which is the graphical sum of E_1, E_2, E_3, E_4 , and E_5) approaches $5E_1$ for very small values of ϕ , E_r must be proportional to $1/\lambda$; hence the field of a Huygens source is proportional to $1/\lambda$, another very useful conclusion.

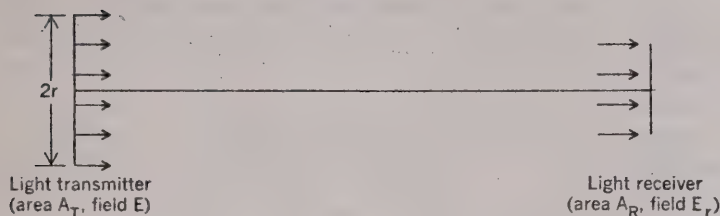


FIGURE 14. Propagation of light between a light transmitter and a light receiver.

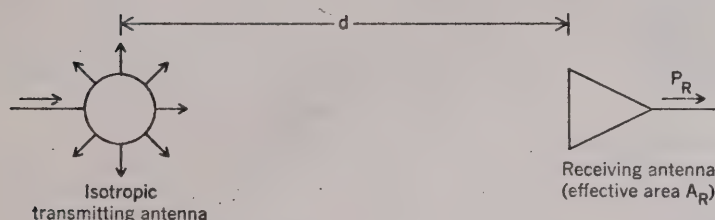
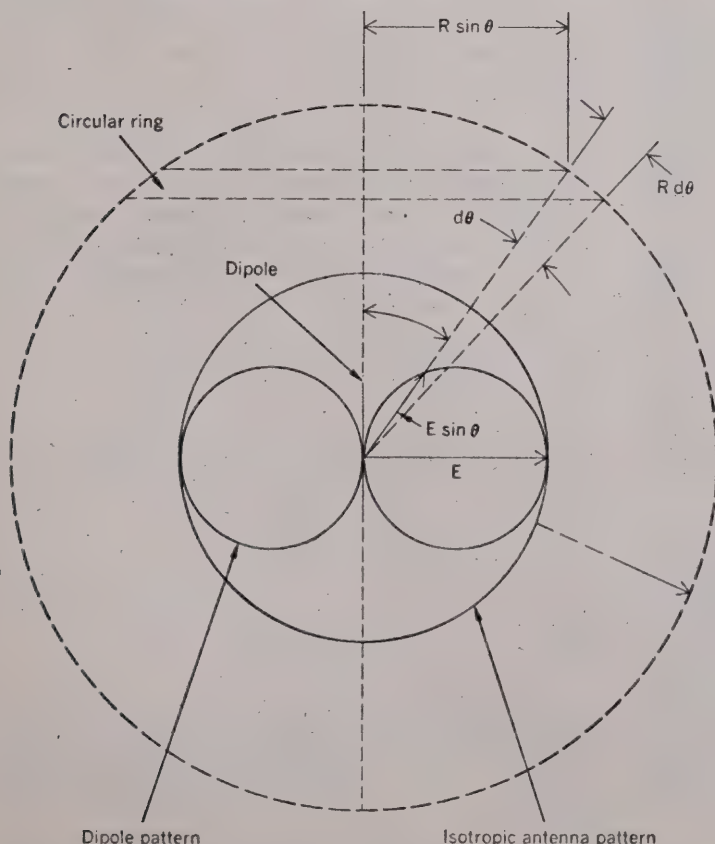


FIGURE 15. Relation between the gain and the effective area of an antenna.

FIGURE 16. Gain of a small dipole.



area A_R and the power transmitted by a light transmitter with area $A_T = \pi r^2$. Figure 14 shows the light circuit. The received and transmitted power flows per unit area are proportional to the square of the light voltages. Hence,

$$\frac{P_R}{P_T} = \left(\frac{E_R}{E} \right)^2 \frac{A_R}{A_T} = \left(\frac{\pi r^2}{\lambda d} \right)^2 \frac{A_R}{A_T} = \frac{A_T A_R}{\lambda^2 d^2} \quad (15)$$

Since light waves and radio waves obey the same laws, A_T and A_R may be considered to be the effective areas of radio transmitting and receiving antennas; this formula is identical to my transmission formula for a radio circuit¹ and it is interesting that it could have been derived long before Maxwell published his theory of electromagnetic waves in 1864 and Hertz discovered radio waves in 1888. Also, the formula applies to acoustic waves. Hence, there should be a derivation independent of electromagnetic theory.

Relation between the gain and effective area of an antenna. Figure 15 shows a hypothetical isotropic antenna that radiates a uniform field in all directions. The power flow at distance d is equal to the transmitted power P_T divided by $4\pi d^2$, the area of a surrounding sphere, and the received power is

$$P_R = P_T \times \frac{1}{4\pi d^2} \times A_R \text{ watts}$$

Replacing the isotropic antenna with a transmitting antenna of area A_T and using Eq. (15), we have, for the received power,

$$P_R' = P_T \frac{A_T A_R}{d^2 \lambda^2} \text{ watts}$$

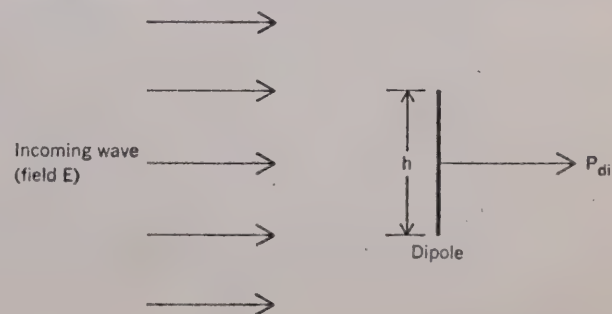
The gain g of the new transmitting antenna, defined as P_R'/P_R , is

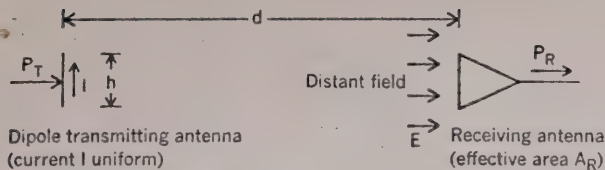
$$g = 4\pi \frac{A_T}{\lambda^2} \quad (16)$$

and this important formula applies to receiving antennas as well.

The gain of a small dipole. A wire carrying an alternating current radiates power. Hertz showed this experimentally years ago and later work showed that the radiation field in a direction is proportional to the projected length of the wire in that direction. Figure 16 shows the radiation patterns of a current element or small dipole (cross section of two circles with radii $E/2$) and an iso-

FIGURE 17. Effective area and radiation resistance of a small dipole.





Dipole transmitting antenna (current I uniform)

FIGURE 18. The distant field E from a small dipole with uniform current I .

tropic antenna (cross section of a circle with radius E). E is the field at a large distance R . The broken-line circle with radius R is the cross section of a sphere. The surface of the sphere is divided into circular rings with area $2\pi R \sin \theta \times R d\theta$. The power flow through the ring due to the dipole is

$$dP_{\text{dipole}} = \frac{\frac{1}{2} E^2 \sin^2 \theta}{120\pi} \times 2\pi R \sin \theta \times R d\theta$$

$$= \frac{\frac{1}{2} E^2}{120\pi} \times 2\pi R^2 \sin^3 \theta d\theta$$

The power radiated by the dipole is then

$$P_{\text{dipole}} = \frac{\frac{1}{2} E^2}{120\pi} 2\pi R^2 \times \int_0^\pi \sin^3 \theta d\theta = \frac{\frac{1}{2} E^2}{120\pi} 2\pi R^2 \times \frac{4}{3}$$

The power radiated by the isotropic antenna is

$$P_{\text{iso}} = \frac{\frac{1}{2} E^2}{120\pi} 4\pi R^2$$

The gain of the dipole is then

$$g_{\text{di}} = \frac{P_{\text{iso}}}{P_{\text{di}}} = 1.5 \quad (17)$$

Effective area of a small dipole (Fig. 17). Equations (16) and (17) give

$$A_{\text{di}} = g_{\text{di}} \times \frac{\lambda^2}{4\pi} = 1.5 \times \frac{\lambda^2}{4\pi} \approx \frac{\lambda}{2} \times \frac{\lambda}{4} \text{ square meters} \quad (18)$$

Note that this area is independent of the dipole length h (which is always small compared with the wavelength).

Radiation resistance of a small dipole. A small dipole with alternating current radiates power and, therefore, it has a resistance called the radiation resistance.

The power received by the dipole shown in Fig. 17 is

$$P_{\text{di}} = \frac{\frac{1}{2} E^2}{120\pi} A_{\text{di}} = \frac{\frac{1}{2} E^2}{120\pi} \times 1.5 \times \frac{\lambda^2}{4\pi} \text{ watts} \quad (19)$$

The dipole is equivalent to a generator with resistance R_{ra} and induced voltages $E \times h$; the available power is

$$P_{\text{di}} = \frac{\frac{1}{2} (Eh)^2}{4R_{\text{ra}}} \text{ watts} \quad (20)$$

Equations (19) and (20) give

$$R_{\text{ra}} = 80\pi^2 \left(\frac{h}{\lambda} \right)^2 \text{ ohms} \quad (21)$$

Note that the radiation resistance is small when h/λ is small, and this makes it difficult to construct an efficient matching circuit to an output line.

The distant field from a small dipole with uniform

current. For the circuit shown in Fig. 18, Eq. (15) gives

$$P_R = P_T \frac{A_{\text{di}} A_R}{d^2 \lambda^2} \text{ watts} \quad (22)$$

We also have

$$P_R = \frac{\frac{1}{2} E^2}{120\pi} A_R \text{ watts} \quad (23)$$

and, since I is maximum value of the alternating current,

$$P_T = \frac{1}{2} R_a I^2 \text{ watts} \quad (24)$$

Equations (18) and (21)–(24) give

$$E = 60\pi \frac{h}{d\lambda} I \text{ volts per meter}$$

Summary

Assuming unobstructed free-space transmission and using for the most part only mathematics taught in many high schools, it has been possible to derive the fundamental properties of radio antennas and radio transmission.

Dr. J. R. Pierce got me started on this article in the fall of 1969 by encouraging his co-workers to find a simple derivation of my old transmission formula. One of these co-workers, C. L. Ruthroff, mentioned the old work on light waves by Huygens and Fresnel and this set me off in the right direction. Dr. S. A. Schelkunoff suggested that I simplify the discussion by deleting the complex algebra in the original version. Lou Mitchel, physics teacher at Rumson High School, read the paper with one of his students and recommended its publication.

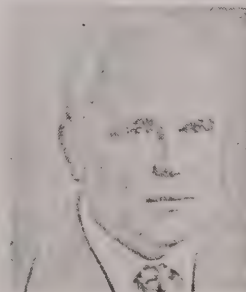
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Harald Trap Friis (F) was born in Naestved, Denmark, in 1893. He received the electrical engineer and doctor of science degrees from the Royal Technical College, Copenhagen, in 1916 and 1933, respectively. In 1919 he moved to the United States and, after a period of study at Columbia University, he joined the Western Electric Company's Research Department, which was later to become Bell Telephone Laboratories, Inc. He was made director of radio research in 1945. In 1952 he became director of research in high frequency and electronics. During his career with the Bell System he contributed substantially to almost every aspect of the radio art, including vacuum tubes, the design of the first commercial superheterodyne receiver, noise, antennas and propagation, radar, and microwaves. Following his retirement from Bell Laboratories in 1958, he served until 1968 as consultant to the Hewlett-Packard Company. At present he is doing consulting work from his home at Rumson, N.J.

Dr. Friis is a member of the American Section, International Scientific Radio Union; and the Danish Academy of Technical Sciences. He was awarded the IRE's Morris Liebmann Memorial Prize (1939) and Medal of Honor (1955), the Franklin Institute's Stuart Balentine Medal (1958), and the IEEE's Mervin J. Kelly Award (1964). He received the Danish decoration "Knight of the Order of Dannebrog," presented by King Frederick IX, in 1954, and the Valdemar Poulsen Gold Medal, presented by the Danish Academy of Technical Sciences, also in 1954.



Developments in broadband antennas

A survey is presented for the purpose of providing the nonspecialist with a basic understanding of the remarkable advances that have taken place over the past decade in the field of broadband antennas

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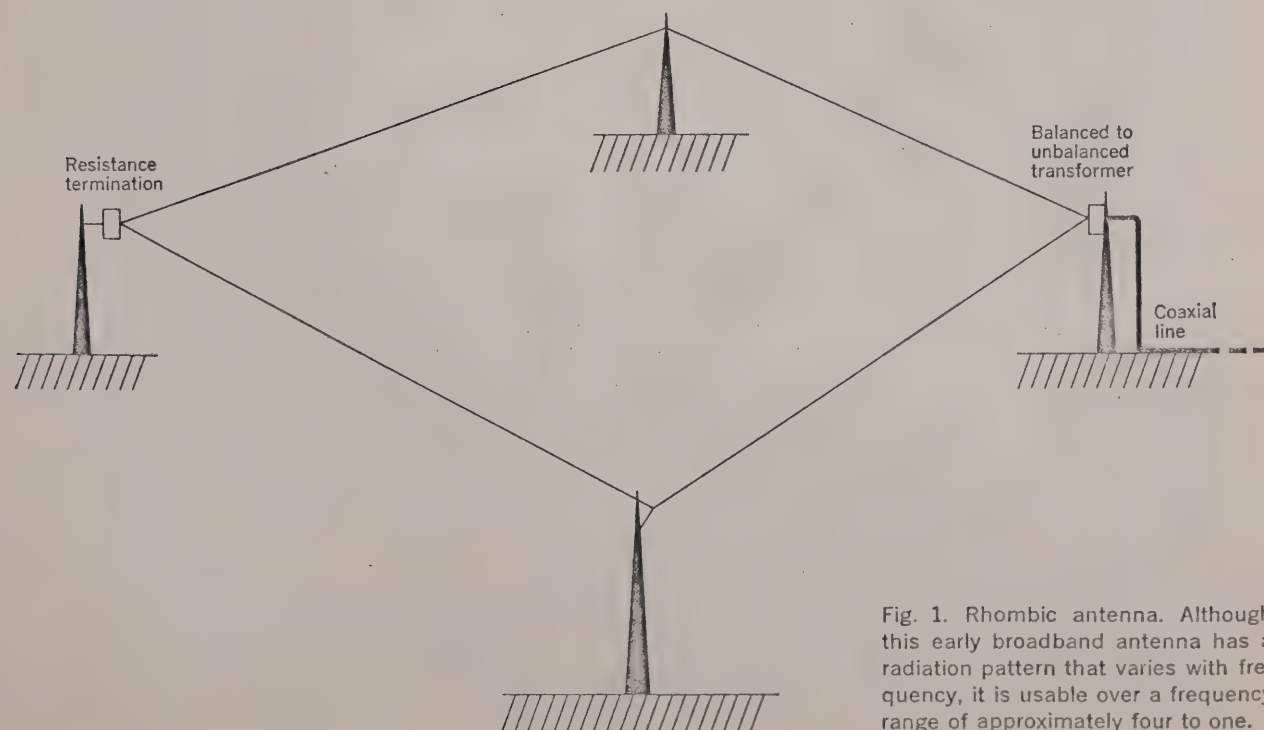


Fig. 1. Rhombic antenna. Although this early broadband antenna has a radiation pattern that varies with frequency, it is usable over a frequency range of approximately four to one.

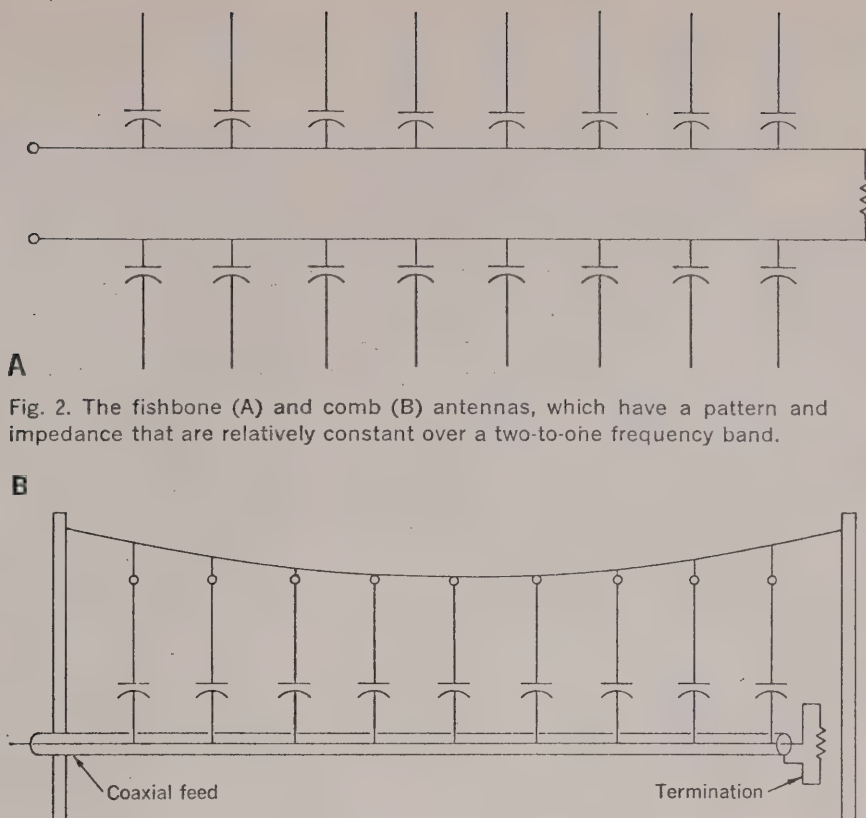


Fig. 2. The fishbone (A) and comb (B) antennas, which have a pattern and impedance that are relatively constant over a two-to-one frequency band.

Until a few years ago the ultimate limiting factor on the bandwidth of any communication system using radiated-wave propagation was most frequently the antenna. Since the antenna performs the dual functions of an impedance-matching device and a directional radiator, the characteristics of major importance are its impedance and its directional properties. Depending upon the application, one or the other (or both) of these characteristics may limit the useful bandwidth. The required or desired bandwidth also varies markedly with application; for example, the required bandwidth may vary from a few cycles per second for a single-channel VLF system up to about 6 Mc/s for a single television channel. For multichannel operation, the radio broadcast band covers a three-to-one bandwidth, the VHF television band covers a four-to-one bandwidth, and the high-frequency communication band covers a ten-to-one bandwidth, from 3 to 30 Mc/s. Finally, for countermeasures work it is usual to state that the desired frequency range extends from "dc to light."

Because of the wide variety of operational requirements that exist, there is no unique definition of antenna bandwidth. For our purposes, a broadband antenna will be considered to be one which retains certain desired or specified radiation pattern, polarization, or impedance characteristics over more than an octave (that is, a two-to-one frequency range).

Early broadband antennas

Some of the earliest broadband antennas were long-wire types designed to operate in the high-frequency (short-wave) band or in the low-frequency band. For the most part, they were broadband only in the sense that impedance remained relatively constant over the useful

range; in general, no attempt was made to achieve a constant pattern. Among these antennas the well-known rhombic antenna has held a dominant place since the early days of radio. This antenna (Fig. 1) is essentially a resistance-terminated transmission line that has been opened out to form the four sides of a rhombus. Because of the traveling-wave current distribution along the terminated line, the main beam is in the forward direction (toward the termination) at an elevation angle that depends, in a complicated fashion, on the included angle of the rhombus and the lengths of the sides in wavelengths. Fortunately the beam is quite broad in the vertical plane and the angle above ground of the maximum increases as the frequency decreases. This change of angle with frequency is in the correct direction for transmission or reception of ionospherically reflected waves, so a rhombic antenna of fixed dimensions is usable over a wide frequency range (of the order of four to one) in the short-wave band.

The wave antenna, consisting of a long, elevated wire parallel to the ground and resistance-terminated at both ends, is another traveling-wave-type antenna. In contrast with most antennas, which operate best over a highly conducting ground, the wave antenna depends for its operation upon the finite conductivity of the earth beneath it. An incident radio wave traveling along the surface of a finitely conducting earth has a forward tilt and a horizontal component of electric field intensity. It is this horizontal component of electric field, produced by the finite earth conductivity, that induces a traveling wave of voltage in the horizontal wire and the resulting antenna action. Because the antenna has an impedance that is nearly independent of frequency, it is known as an aperiodic antenna. However, the radiation pattern

does vary with the length of the wire in wavelengths, and hence with frequency. Wave antennas are used for long-wave or low-frequency reception.

The fishbone receiving antenna consists of a long resistance-terminated transmission line loosely coupled by capacitors to an array of closely spaced (less than $\lambda/4$), untuned, horizontal dipoles; see Fig. 2(A). For vertically polarized signals, one half of a fishbone antenna is erected vertically and fed against ground to form a comb antenna; see Fig. 2(B). Tapering the coupling capacitors to larger values towards the termination equalizes the antenna currents and reduces resonance effects. Because the capacitive coupling of the elements to the transmission line is tighter at the higher frequencies, fewer of the elements are strongly excited. Hence, the effective length of the array varies inversely with frequency in such a manner as to maintain a fairly constant pattern, gain, and impedance, over the useful bandwidth of more than two to one. The fishbone and comb antennas have been described chiefly for comparison with the log-periodic dipole and log-periodic monopole types to be described later.

In contrast to the terminated wire and loaded transmission line types just described, there is a class of antennas that owes its broadband properties to broad, specially shaped surfaces. It was recognized quite early that a fat dipole had a much lower antiresonant (full-wave-length) impedance than a thin one, and that in general, fat antennas had smaller impedance variations than thin ones. The importance of broad surfaces was emphasized by Schelkunoff in the treatment of the biconical antenna, and many broad-surfaced specially shaped antennas found early application in television transmitting antennas and countermeasures antennas. A very successful broadband antenna was the discone¹ (a cone fed against a disk), which maintained good impedance and pattern characteristics over a four-to-one bandwidth (Fig. 3). For countermeasures work many antennas having surfaces of various shapes were developed,² some of which had remarkably wide impedance bandwidths and usable pattern bandwidths of the order of five to one. It must be admitted, however, that most of these early designs were arrived at by an intuitive or cut-and-try approach.

Another group of antennas, some of which display fairly wide bandwidths, consists of various helical and spiral shapes. When the circumference of a helical antenna is of the order of a free-space wavelength, the antenna radiates in the axial mode—that is, with the maximum radiation along the axis of the helix. In this mode, the helical antenna has desirable impedance, pattern, and circular polarization properties over nearly an octave.³ By expanding the diameter of the helix along its length to form a conical monofilar helix fed from the base end, Springer⁴ showed that the bandwidth could be increased. His observation that there appeared to be an effective aperture that moved toward the smaller end of the cone as the operating wavelength decreased was perhaps the first indication of things to come. Later, Chatterjee^{5,6} also considered monofilar helical antennas formed on a conical surface and fed against a ground plane. He demonstrated that they could be excited from either end, and obtained usable bandwidths of approximately four to one. At about the same time, Turner⁷ proposed a balanced antenna constructed in the form

of an Archimedes spiral. This planar antenna, constructed with narrow constant-width arms and radiating a broad lobe on each side of the structure, gave promise of being usable over the then remarkable bandwidths of between seven and eight to one.

Frequency-independent antennas

In 1954, Rumsey⁸ put forth the idea that a structure entirely definable by angles, without any characteristic length dimension, should have properties that are independent of the frequency of operation. However, all such angle structures extend to infinity, so the key question was which of such structures retained these frequency-independent characteristics when truncated to a finite length. It should be noted that the well-known biconical structure is an angle structure that is *not* frequency independent when it is truncated to form a practical antenna. Both impedance and pattern vary with frequency for any finite length.

Rumsey proposed that an equiangular spiral structure, which satisfies the angle requirement, might have the desired properties, and Dyson⁹⁻¹¹ undertook a comprehensive experimental study of an antenna based on the equiangular spiral geometry shown in Fig. 4. The equiangular or logarithmic spiral* is defined by

$$\rho = e^{a(\phi - \delta)} \quad \text{or} \quad \phi - \delta = \frac{1}{a} \ln \rho$$

where ρ and ϕ are conventional polar coordinates, and a and δ are constants. In Fig. 4 the edges of the metallic arms are defined by

$$\rho_1 = ke^{a\phi} \quad \text{and} \quad \rho_2 = ke^{a(\phi - \delta)}$$

for one arm, and by

$$\rho_3 = ke^{a(\phi - \pi)} \quad \text{and} \quad \rho_4 = ke^{a(\phi - \pi - \delta)}$$

for the other arm, where the constants a , k , and δ determine the rate of spiral, size of the terminal region, and arm width, respectively. With this particular spiral the angle between the radius vector and the spiral remains the same for all points on the curve—hence the term “equiangular spiral.” Experimental investigation established that this particular geometry did indeed retain its frequency-independent properties after truncation, and this design was the basis for a large class of successful frequency-independent antennas.

When this angular structure is excited in a balanced manner at the origin, the current flows outward with small attenuation along the spiral arms until a region of given size in wavelengths is reached. In this region (the active or radiating region) essentially all of the incident energy transmitted along the spiral arms is radiated, and somewhat beyond this region the presence or absence of the arms is of no consequence. Because the radiating region is of constant size in wavelengths, it moves toward the origin as the wavelength of operation decreases. The size of effective radiating aperture thus automatically adjusts or scales with frequency of operation in such a manner that the antenna behaves the

* The logarithmic spiral was first discussed by Descartes (1638) and later (1691–1693) studied by Jacques Bernoulli, who gave it its name. Bernoulli was so delighted by the property of the spiral reproducing itself under various transformations that he requested that the spiral be engraved on his tomb with the inscription “Eadem Mutata Resurgo.”

same at all frequencies. Because of the spiraling of the arms, this scaling is accompanied by a rotation of the radiated field about the axis of the antenna.

It is now known that this automatic scaling of the radiating aperture is a condition for operation in a frequency-independent manner. It is interesting to note that Springer observed this phenomenon on the expanding helix, but unfortunately the methods of construction and excitation limited the bandwidth obtainable to something over an octave, so the importance of scaling of effective aperture with frequency was not fully recognized. Chatterjee's measurements also show evidence of scaling with frequency in the near-field amplitude plots from which he calculated radiation patterns; but again, possibly because of the physical configuration and method of feed, the full significance of this scaling does not appear to have been appreciated. In a similar manner, the radiating aperture of the Archimedes spiral antenna tends to scale with frequency; however, because the width and spacing of the spiral arms in the radiating region are not constant in wavelengths, as frequency is varied, the antenna characteristics change (albeit slowly) with frequency.

At this point it is necessary to define the term "frequency independent" when it is used with a practical finite-sized structure. If the antenna is excited by a voltage applied between the two arms at the origin, it has an impedance and radiation pattern that are essentially constant* (that is, independent of frequency) for all frequencies above that for which the outer diameter of the truncated structure is approximately half a wavelength up to the frequency at which the diameter of the feed region (as determined by the transmission line feed) is comparable with a half wavelength. Since these two dimensions can be specified independently, the design bandwidth can be made arbitrarily large; actually it is limited only by practical considerations of construction—that is, how large the outer diameter is made and how finely the geometry at the feed region can be modeled.

The equiangular spiral antenna, which is bidirectional, radiates a very broad, circularly polarized beam on both sides of its surface. This bidirectional characteristic severely restricts its utility in practice, but a modified version, to be described later, provides a highly practical, extremely broadband antenna.

Log-periodic antennas

In 1955, working with Rumsey on broadband antenna development, DuHamel¹² proposed that it should be possible to force radiation from otherwise "angle structures" by the use of appropriately located discontinuities. One of the first geometries chosen to investigate the validity of this concept was that shown in Fig. 5. Here two wedge-shaped metallic angle structures have teeth cut into them along circular arcs. The radii of the arcs which define the location of successive teeth are chosen to have a constant ratio $\tau = R_{n+1}/R_n$. This same ratio τ defines the lengths and the widths of suc-

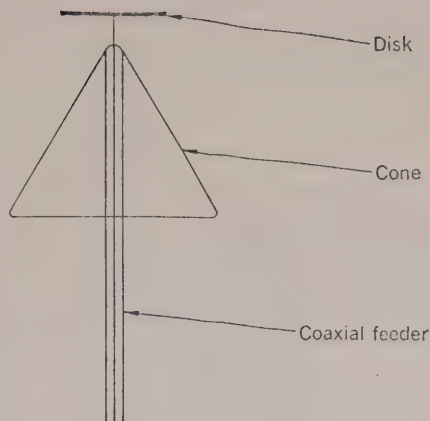


Fig. 3. The discone, a successful early broadband antenna having a useful bandwidth of approximately four to one.

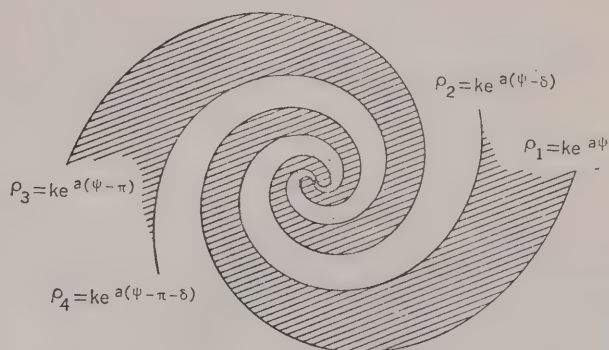
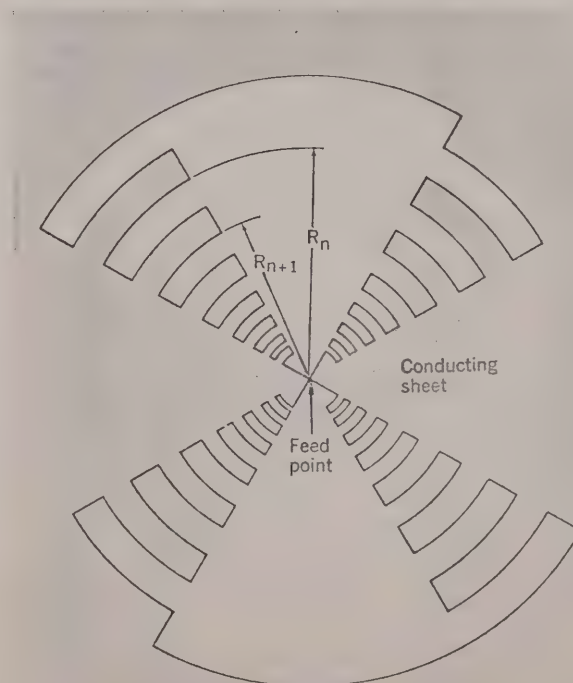


Fig. 4. Sketch showing the geometry of the equiangular (or logarithmic) spiral antenna with equations of the edges.

Fig. 5. A sheet-metal log-periodic antenna.



* The pattern actually rotates with frequency about an axis perpendicular to the plane of the spiral. If the pattern-measuring coordinate system is allowed to rotate at the same rate, the measured pattern remains constant; otherwise there will be a (generally small) periodic variation of magnitude proportional to the rotational asymmetry of the pattern.

cessive teeth. From the principle of modeling it is evident for this structure, extending from zero to infinity and energized at the vertex, that whatever properties it may have at a frequency f will be repeated at all frequencies given by $\tau^n f$, where n is an integer. When plotted on a logarithmic scale, these frequencies are equally spaced with a period equal to the logarithm of τ ; hence the name "log-periodic" structure. Log-periodicity guarantees only periodically repeating radiation pattern and impedance. However, for certain types of such structures and for values of τ not too far from unity, variation of characteristics over a period can be quite small, and an essentially frequency-independent structure results. It is important to note that only a relatively few of the nearly infinite variety of log-periodic structures will make successful broadband antennas in the sense that the impedance and pattern characteristics will remain constant when the structure is truncated to a finite length. It happens that the geometry of Fig. 5 did result in a successful log-periodic antenna.

The antenna of Fig. 5 was designed to have one other rather special property; namely, that the metal cut away from the plane sheet to form the antenna arms has identical shape with the metal that remains. In other words, the complementary slot antenna has the same size and shape as the metallic dipole antenna. Now by an extension of Babinet's principle it is known that complementary-dipole and slot antennas have impedances Z_d and Z_s , respectively, related by $Z_d Z_s = (60\pi)^2$. Because the slot antenna and dipole antenna are the same (for the geometries chosen) it follows that $Z_d = Z_s = 60\pi \approx 189$ ohms, a result that is independent of frequency. Hence this particular geometry assured constant impedance, although not constant radiation pattern, independently of the other consideration of log-periodic geometry. In view of this use of Babinet's principle in the design of these planar structures, the next step to be taken was a bigger one than might at first appear.

Unidirectional frequency-independent and log-periodic antennas. Both the equiangular spiral antenna (Fig. 4) and the log-periodic antenna (Fig. 5) radiate equally on both sides of the plane of the antennas, a result that severely limits their usefulness. A major step forward was made in extending the range of practical application when Isbell¹³ bent the two arms of the planar log-periodic structure toward each other (out of the plane) to form the nonplanar V-shaped antenna of Fig. 6. Two rather surprising results were observed. As the angle between the two arms of the antenna was decreased from 180° the radiation pattern changed from bidirectional to unidirectional, with the major radiation off the apex of the antenna—that is, in the backward direction. Moreover, although one of the necessary conditions for Babinet's principle (that of a plane surface) was now violated, the impedance continued to remain nearly constant with frequency, but at a different value, which depended upon the angle between the arms. This nonplanar version of the log-periodic structure, radiating a plane-polarized unidirectional beam, greatly increased the utility of the log-periodic structures.

The frequency-independent logarithmic spiral structure also found wider use when Dyson developed a unidirectional version by wrapping the balanced spiral arms on the surface of a cone, as shown in the antenna of Fig.

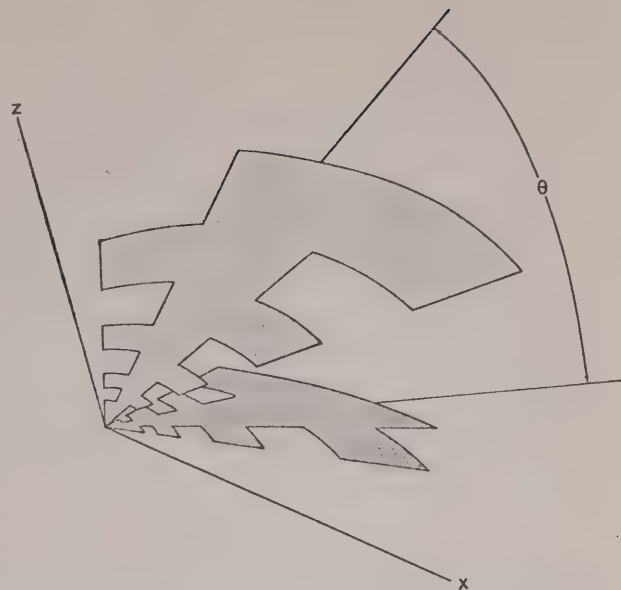
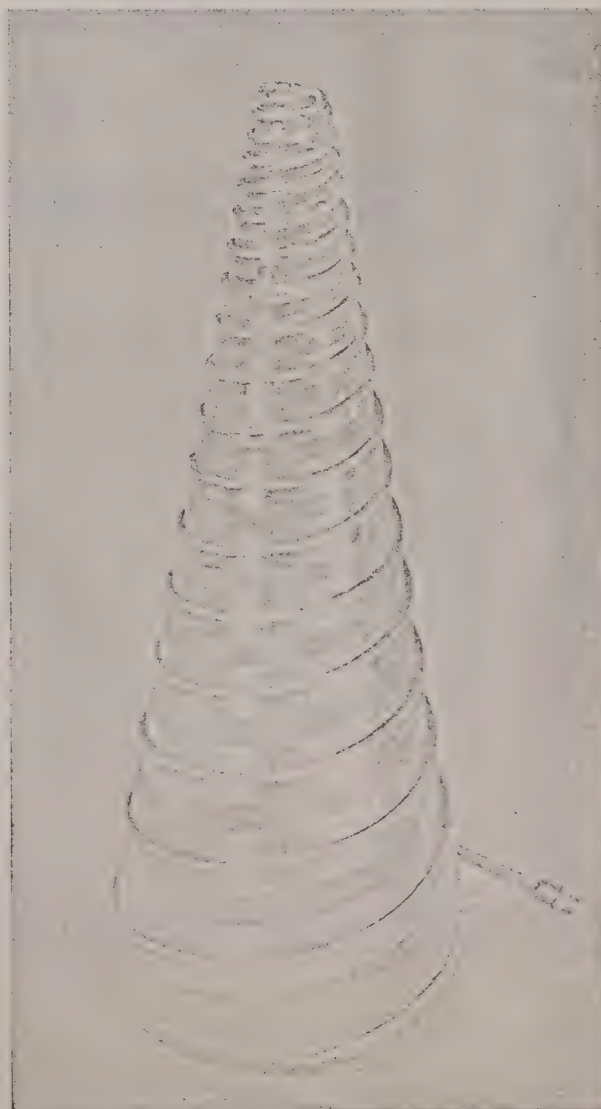


Fig. 6. Nonplanar, unidirectional log-periodic antenna.

Fig. 7. Unidirectional conical spiral antenna.



7. For appropriately chosen rates of spiral this modified version continued to yield essentially frequency-independent performance. For cone angles of less than about 45° the pattern became unidirectional with a broad-lobed beam, again in the backward direction off the apex of the cone.

The conical equiangular spiral antenna is a balanced structure, which may be fed (at the apex) by means of a balanced transmission carried up inside and along the axis of the cone. Alternatively, it may be fed as illustrated in Fig. 7 by a coaxial cable carried along and soldered in contact with one of the arms. Because the amplitude of antenna current on the arms, and also on the outside of the coaxial cable, falls off quite rapidly with distance from the apex, the ends of the arms where the cable enters is essentially a field-free region. This type of feed automatically provides a frequency-independent balun (balanced converter), permitting the balanced antenna to be fed by means of an unbalanced coaxial line. To maintain physical symmetry a dummy cable is usually soldered to the other arm. Conical equiangular or log-spiral antennas have been constructed to operate over bandwidths of higher than 40 to 1. The bandwidth obtained is at the discretion of the designer. The upper usable frequency is determined by the truncated region at the apex, which must remain small in terms of wavelengths, and the lowest usable frequency is set by the base diameter of the cone, which must be at least $\frac{3}{8}$ wavelength at the lowest frequency of operation for spirals that are wrapped fairly tightly.

A further modification of the conical equiangular spiral results in a very practical, easily constructed antenna. If the width of the expanding arms is narrowed and they are allowed to degenerate to constant-width structures, the cables alone can form the arms. For fairly tightly spiraled antennas there is little change in the characteristics from those of an antenna with narrow expanding arms.

Other types of log-periodic antennas. The practical value of the log-periodic approach was enhanced even

further when DuHamel¹⁴ and co-workers demonstrated that successful log-periodic antennas could be made with wire structures as well as sheet structures. This development extended the range of application down from microwaves through the high-frequency band. A typical wire version of a log-periodic antenna is shown in Fig. 8. It was also demonstrated that for higher gain a frequency-independent array of log-periodic antennas could be constructed by arranging the antennas like the spokes of a wheel with the origins of the individual antennas at the hub.

Still another application of the log-periodic principle is the log-periodic dipole array¹⁵ of Fig. 9. As with all log-periodic geometries, all dimensions are increased by a constant ratio in moving outward from the origin. Thus the lengths and spacings of adjacent elements must be related by a constant scale factor τ , as follows:

$$\frac{l_n}{l_{n-1}} = \frac{d_n}{d_{n-1}} = \tau$$

Although at first glance this antenna might appear similar to the early fishbone antenna with $\tau = 1$, there are several essential differences. For successful operation, the log-periodic dipole array must be fed with a transposition of the transmission line between adjacent dipole elements. The antenna is then caused to radiate in the backfire direction (that is, toward the source), a condition which appears to be necessary for successful unidirectional frequency-independent or log-periodic operation. The

Fig. 8. Log-periodic wire antenna for frequencies of 11 to 60 Mc/s. (Photo courtesy Collins Radio Company.)

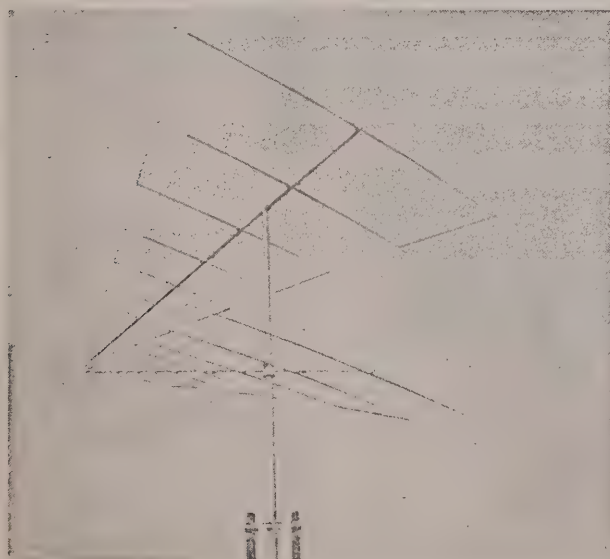
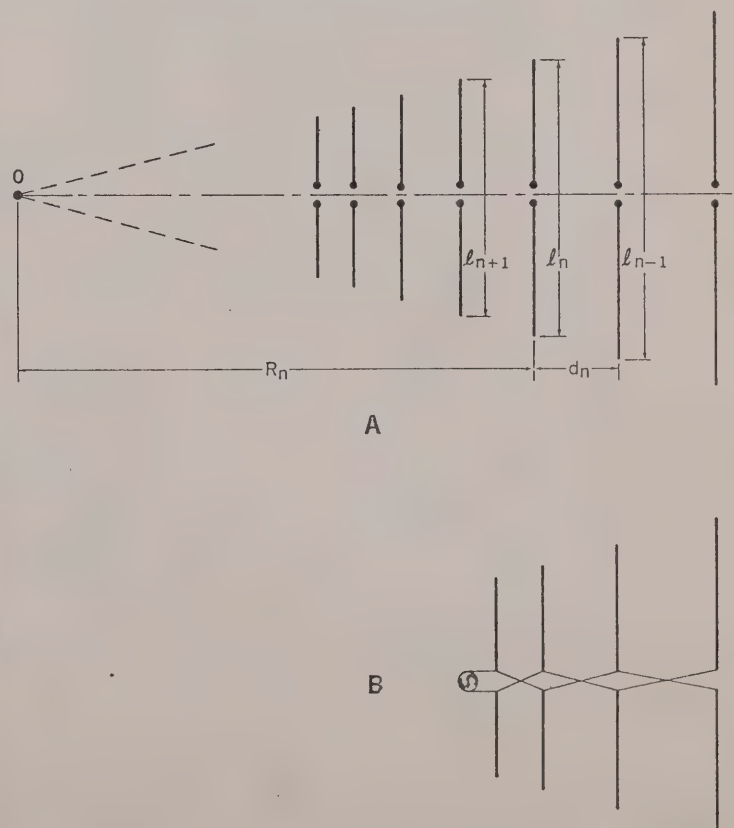


Fig. 9. Log-periodic dipole antenna array. A—Lengths and spacings of elements. B—Method of feeding.



active portion of the array from which most radiation occurs is centered around those elements near resonance (for which I_n is somewhat less than $\lambda/2$). As the frequency is changed the active region moves back or forth along the array. Because practically all of the input power is absorbed in and radiated by the active portion, the larger elements to the right of the active region are not excited. Moreover, because the beam is directed toward the feed point at the left, these larger elements are in an essentially field-free region, and so do not adversely affect the operation. The shorter elements to the left of the active region are in the beam but, because of their short lengths, close spacings, and alternate phasings, have small influence on the pattern.

Basic principles of operation of log-periodic and frequency-independent antennas. Of the almost unlimited variety of log-periodic structures that can be devised, only a small fraction will produce successful antennas

when truncated. It is interesting to search out the essential requirements for successful design. The operation of the log-periodic dipole array of Fig. 9, being simple and easily understood, will be analyzed in some detail. From the understanding so gained it should be possible to extend the analysis to less familiar geometries, and then to frequency-independent antennas in general.

At this point it will be advantageous to recall some of the basic notions of antenna array theory. Consider an n -element array of equispaced isotropic radiators (Fig. 10) having equal current amplitudes and a spacing d less than one-half wavelength. (An isotropic radiator is one that radiates uniformly in all directions; a simple dipole antenna is an isotropic radiator in the H plane perpendicular to its axis.) At a distant point the electric fields from these radiators will add with a phase angle between them which is dependent upon the relative phasings of the radiator currents and the relative phase delays produced by the difference in path lengths to the distant point. For the array shown the phase difference due to path length difference between adjacent elements is $(2\pi/\lambda)d \cos \phi$ radians. If the elements of the array are fed with a progressive phasing of currents equal to α , where α represents the angle by which the current in a given element *leads* the current in the preceding element, then at the distant receiving point the phase difference of the fields produced by adjacent elements will be

$$\psi = \alpha + \frac{2\pi}{\lambda} d \cos \phi = \alpha + kd \cos \phi \quad (1)$$

where $k = 2\pi/\lambda$ is the free-space phase-shift constant. The total electric field at any distant point will be given by the phasor sum

$$E_t = E_0 1 + e^{j\psi} + e^{j2\psi} + \dots + e^{j(n-1)\psi} \quad (2)$$

where E_0 is the field intensity at the reception point produced by current I_0 . E_t can be obtained graphically

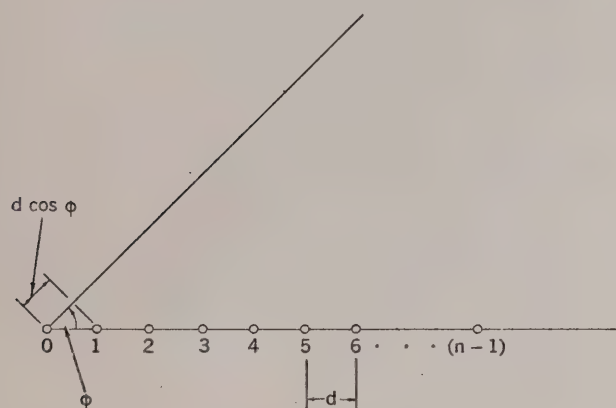
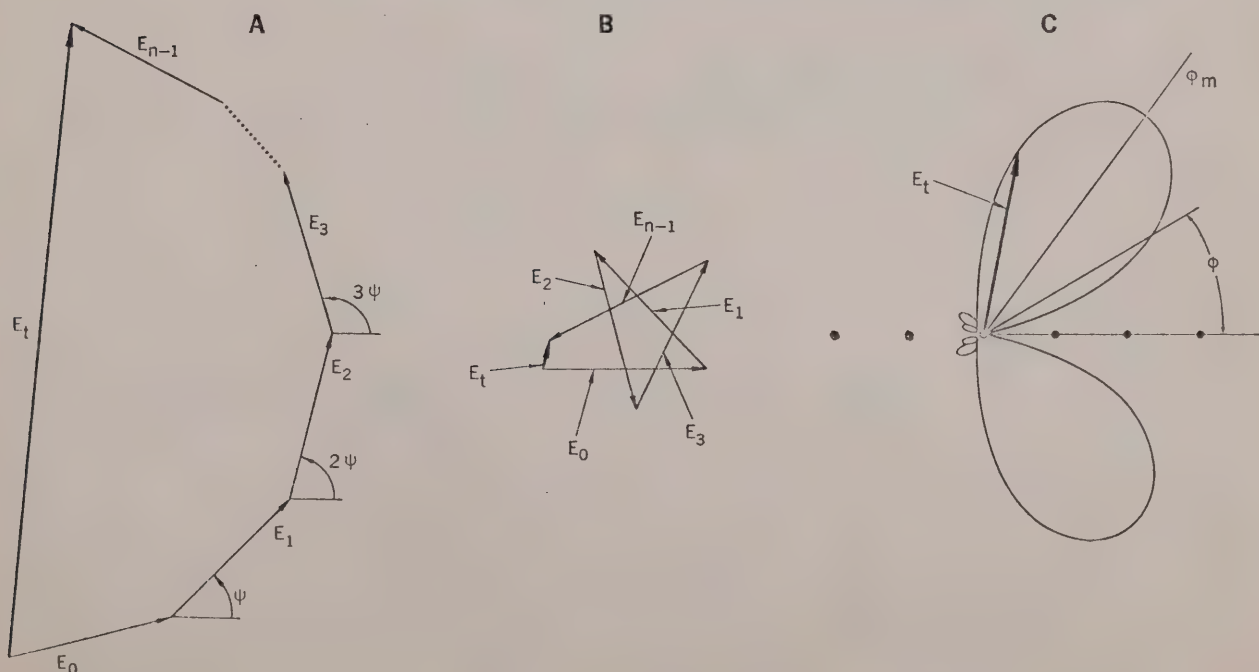


Fig. 10. Array of equispaced isotropic radiators.

Fig. 11. A and B—Graphical construction for E_t . C—Calculated radiation pattern, E_t vs. ϕ .



from the construction of Fig. 11(A). Using the particular value of α , and computing ψ from (1) for various values of ϕ , the construction of Fig. 11(A) can be used to determine a radiation pattern of the array; see Fig. 11(C). It is evident that the total field intensity will be maximum when $\psi = 0$, so that all fields add in phase.

Therefore, for a maximum, $\psi = \alpha + kd \cos \phi = 0$. The angle ϕ_m for maximum radiation is given by

$$\cos \phi_m = -\frac{\alpha}{kd} \quad \text{or} \quad \phi_m = \cos^{-1} \frac{-\alpha}{kd} \quad (3)$$

If the elements are fed in phase, $\alpha = 0$, and $\phi_m = 90^\circ$, so the maximum radiation is broadside. If successive elements are fed with a lagging phase of value, $\alpha = -kd$, then $\phi_m = 0$, so the maximum radiation is endfire in the forward direction. If successive elements are fed with a leading phase of value, $\alpha = +kd$, then ϕ_m will equal 180° , and the maximum radiation will be endfire in the backward direction. For values of α between $-kd$ and $+kd$, the angle of maximum radiation is at an angle between 0 and 180° as given by Eq. (3). By symmetry about the axis of the array, there is another maximum at an angle between 0 and -180° , which is also given by (3). When $|\alpha| > kd$, Eq. (3) cannot be satisfied for any real value of ϕ ; that is, there is no value of ϕ in the "visible" range between 0 and 180° (hence, also between 180° and 360°) that will produce a maximum—in the sense that all the radiations add in phase. However, if $|\alpha|$ is only slightly greater than kd , so that ψ is not much larger than zero, the total field can still be quite strong in the forward direction ($\phi = 0$) for negative α , or in the backfire direction ($\phi = 180^\circ$) for positive α . This case is illustrated by the sketch of Fig. 11(A). On the other

hand, if $|\alpha|$ is considerably greater than kd (that is, the phase shift between elements is large), the phase diagram might be as illustrated in Fig. 11(B), with a resulting small total E_t for all values of ϕ .

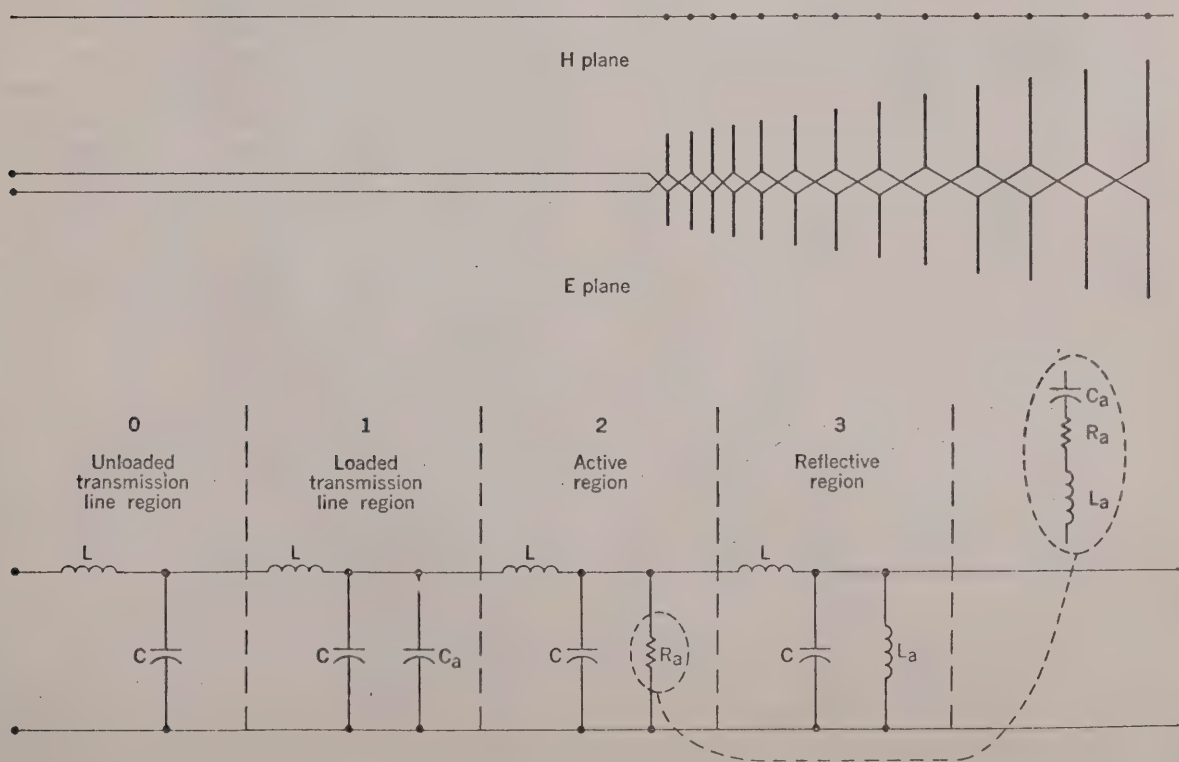
For these cases of large phase shift, as shown in Fig. 11(B), there is no major lobe anywhere, and the array radiates only feebly, scattering its small radiated energy in various directions.

The elementary notions just discussed can be applied with some slight modification to an analysis of the log-periodic dipole array sketched in Fig. 12. For this purpose, it is helpful to consider separately three main regions of the array.

1. Transmission-line region. The antenna elements in the transmission-line region are short compared with the resonant length (that is, $l \ll \lambda/2$), so the element presents a relatively high capacitive impedance. The element current is small and leads the base voltage supplied by the transmission line by approximately 90° . The element spacing is small in wavelengths and the phase reversal introduced by transposition of the transmission line means that adjacent elements are nearly 180° out of phase. More precisely, each element current leads the preceding element current approximately by $\alpha = \pi - \beta d$, where d is the element separation and $\beta = 2\pi/\lambda = \omega/v$ is the phase-shift constant along the line. In general β , λ , and v will differ from their free-space values owing to the loading effect of the elements on the transmission lines. Because of the phasing and close spacing of the elements, radiation from this region will be very small and in the backfire direction.

2. Active region. In the active region the element lengths approach the resonant length (l slightly less than

Fig. 12. Transmission-line representation of log-periodic dipole array.



$\lambda/2$), so the element impedance has an appreciable resistive component. The element current is large and more nearly in phase with the base voltage; the current is slightly leading just below resonance and slightly lagging just above resonance. The element spacing is now sufficiently large to allow the phase of current in a given element to lead that in the preceding element by an angle $\alpha = \pi - \beta d$, which may approximate $\pi/2$ radians. This combination of conditions will produce a strong radiation in the backfire direction.

3. Reflection region. The element lengths in the reflection region are greater than the resonant length ($l \geq \lambda/2$), so the element impedance becomes inductive and the element current lags the base voltage. The base voltage provided by the transmission line is now quite small, because in a properly designed array nearly all of the energy transmitted down the line has been abstracted and radiated by the active region. The element spacing may now be larger than $\lambda/4$. However, as will be shown later, the phase shift per unit length along the line in this region is small, so the resulting phasing between elements (including the phase reversal introduced by the transposition) is such that any small amount of radiation is still in the backfire direction. In addition, it will be demonstrated later that the characteristic impedance of the transmission-line becomes reactive in this region. Thus, any small amount of incident energy transmitted through the active region is not accepted in the reflection region but is reflected back toward the source.

The array as a loaded transmission line. Some of the remarkable properties of log-periodic and frequency-independent antennas are attributable to the propagation characteristics of the equivalent loaded transmission line that conveys energy from the source to the radiating portion of the antenna. These effects are particularly easy to see in the case of the log-periodic dipole array, shown in Fig. 12. On the feed line to the antenna, region 0, the series inductance and shunt capacitance per unit length are shown as L and C , respectively. In the transmission region of the antenna, region 1, the transmission line is loaded by a capacitance per unit length C_a that represents the loading effect of the short dipoles, which have a capacitance reactance. It is noted that to the first approximation C_a is nearly constant throughout this region because at the beginning of the region the capacitance per element is small, but the elements are closely spaced, whereas near the end of the region the capacitance per element is larger, but so is the spacing. The effect of the augmented shunt capacitance of the line ($C + C_a$) is to increase the phase delay per unit length, and since $\beta = 2\pi/\lambda = \omega/v$, this means a decrease of wavelength λ and a decrease of phase velocity v along the line below the free-space values. This is said to be a "slow wave" region of the transmission line. Note, however, that because of the transposition of the feed line between elements, successive elements are fed with a leading phase shift of $\pi - \alpha$ per section. This rapid phase shift in the reverse direction corresponds to a slow wave in the backward direction along the antenna elements.

In region 2, the element lengths approach the resonant length and the transmission line loading becomes resistive, designated by the shunt resistance R_a in series with the antenna capacitance C_a and antenna inductance L_a . The phase shift per unit length, the wavelength, and the phase velocity all approach their free-space values.

Because of the transposition between elements, and accounting for the fact that the element current leads the base voltage by lesser amounts in successive elements as the resonant length is approached, it turns out that phasing of currents in the elements corresponds to a backward traveling wave having a velocity v somewhat less than c , the velocity of light.

In region 3 the element lengths become longer than the resonant length, the antenna inductive reactance pre-

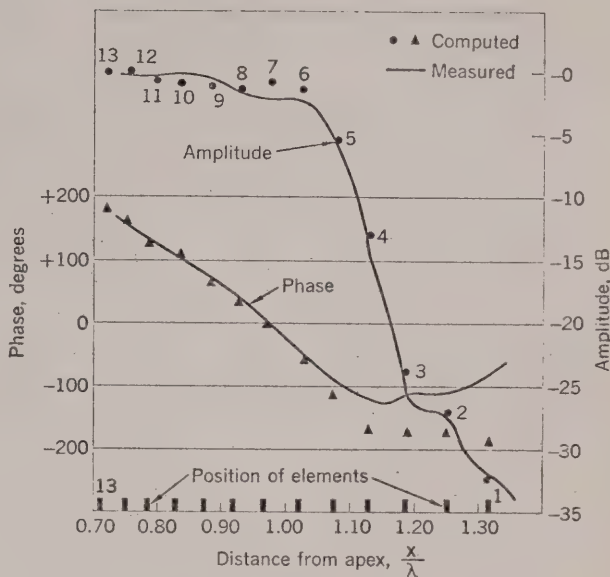
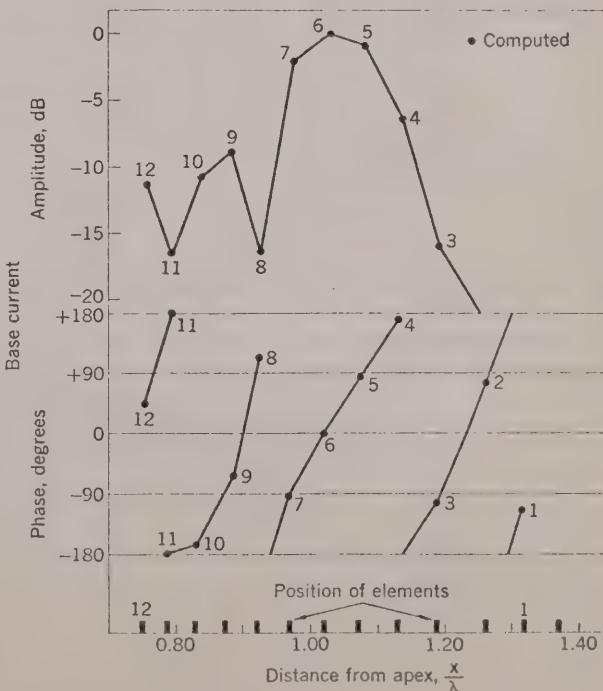


Fig. 13. Base voltages along a typical 13-element log-periodic dipole array at a frequency for which element 4 is half of a wavelength.

Fig. 14. Element currents corresponding to Fig. 13.



dominates, and the loading effect on the line is represented by the shunt inductance L_a . If the parallel combination of L_a and C is inductive, we have the equivalent of the attenuation region of a filter. The phase shift per unit length is then zero (for the lossless case) and the phase velocity is infinite; that is, there is no wave motion. The incident energy propagating down the line is no longer accepted but is reflected back toward the source. These results are strictly true only in the case of a lossless filter, but they form the first approximation in the case of a lossy filter.

The general features outlined in the foregoing discussion will be illustrated for a particular log-periodic dipole array, which has been analyzed in considerable detail.¹⁶ Fig. 13 shows the amplitude and phase of the transmission line voltage along a particular 13-element log-periodic dipole array. Distance is shown measured from the apex of the array, and the elements are numbered starting with the largest element as number 1. This set of data is for a frequency f for which element number 4 is $\lambda/2$ long. Several interesting aspects of the data are immediately apparent: In the transmission region (elements 13 to 7), the amplitude of voltage along the line is approximately constant and the phase shift between element positions increases gradually from about 20° to 30° . (Because of the transposition between elements, this means that adjacent elements are fed with a progressive phase lead 160° to 150° .) In the active region (elements 7 to 4) the amplitude drops sharply because of power absorbed by the strongly radiating elements, and the phase shift averages about 90° between adjacent elements. Finally, in the unexcited or reflection region (elements 3 to 1), the amplitude drops to very low values and the phase shift between element positions is nearly zero (corresponding to the zero phase shift or infinite phase velocity in the attenuation region of a low-pass filter).

The resulting element currents for the log-periodic dipole array of Fig. 13 are shown in Fig. 14, both in amplitude and phase. From the current amplitudes (noting that small contributions from elements 12 through 8 tend to cancel one another because of the nearly 180° phase shift between them), it is evident that the only elements that will contribute appreciably to the radiation are elements 7, 6, 5, and 4. For these elements, the phase difference between adjacent members is approximately 90° leading, so a backfire radiation will be expected. The phasor diagrams for $\phi = 0^\circ, 90^\circ$, and 180° are shown in Fig. 15 and the resulting radiation patterns are shown in Fig. 16. (The E-plane pattern is the H-plane pattern modified by the directivity of the individual elements in this plane.)

As operating frequency is decreased or increased the active region moves up or down the array, but radiation pattern and input impedance remain almost constant.

General properties of log-periodic and frequency-independent antennas. The manner of operation of the log-periodic dipole array has been described in some detail because of the insight it gives into what are believed to be general requirements for successful frequency-independent operation. These appear to be as follows:

1. An excitation of the antenna or array from the high-frequency or small end of the antenna.
2. A backfire radiation (in the case of unidirectional radiators), so that the antenna fires through the small

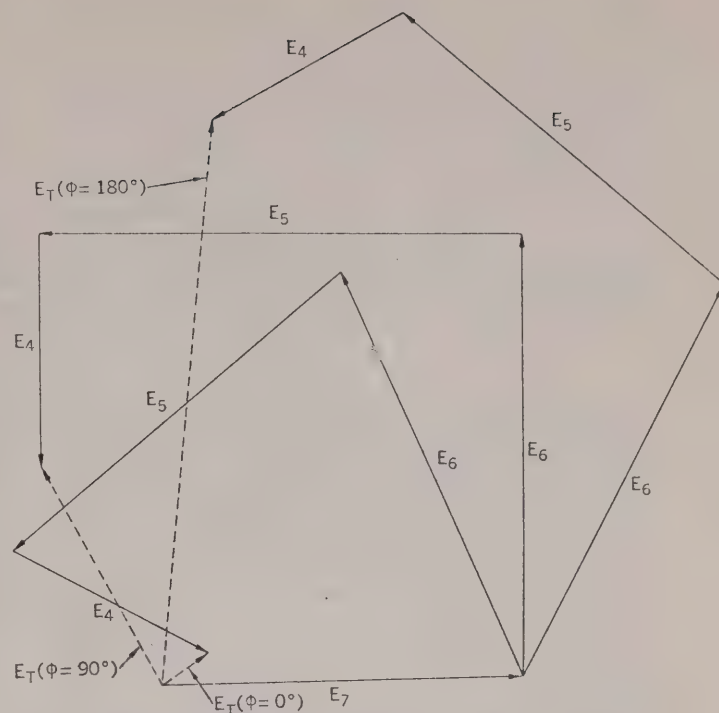
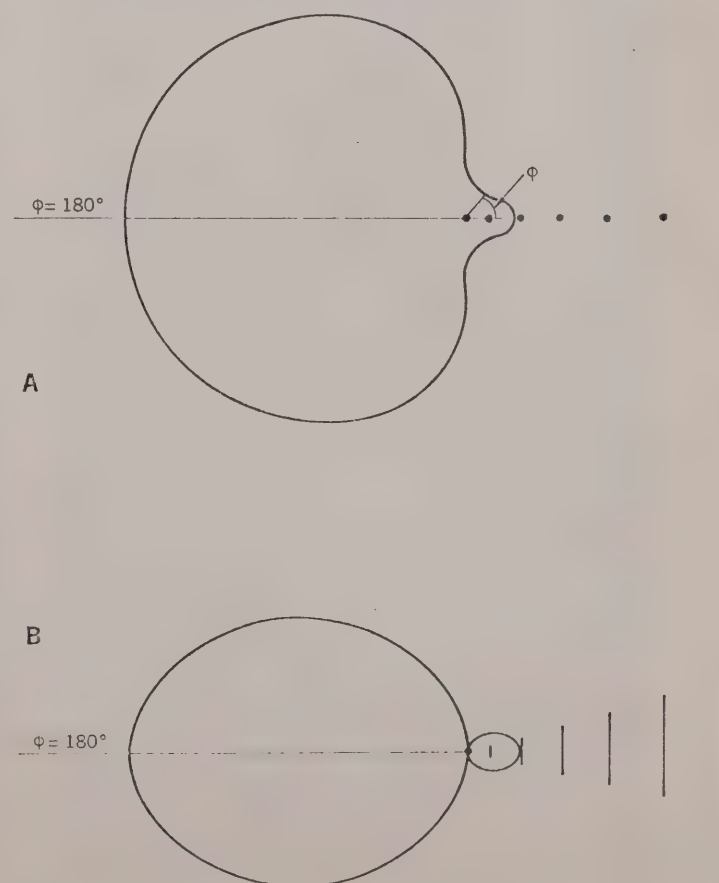


Fig. 15. Phasor sum of radiated fields from currents shown in Fig. 14, for ϕ values of $0^\circ, 90^\circ$, and 180° .

Fig. 16. Radiation patterns resulting from fields shown in Fig. 15. A—H-plane pattern. B—E-plane pattern.



part of the antenna, with the radiation in the forward direction being zero or at least very small. For bidirectional antennas the backfire requirement is replaced by a requirement for broadside radiation. In any case, the radiation in the forward direction along the surface of the antenna (which theoretically extends to infinity) must be zero or very small.

3. A transmission region formed by the inactive portion of the antenna between the feed point and the active region. This transmission line region should have the proper characteristic impedance and negligible radiation.

4. An active region from which the antenna radiates strongly because of a proper combination of current magnitudes and phasings. The position and phasing of these radiating currents are such as to produce a very small radiation field along the surface of the antenna or array in the forward direction, and a maximum radiation field in the backward direction (broadside for bidirectional antennas). For successful backfire antennas these requirements are frequently met with separations less than a quarter wavelength and phasings near 90° leading, for adjacent elements in the active region. For broadside radiation the phasings must, of course, be zero.

5. An inactive or reflection region beyond the active region. All successful frequency-independent antennas must exhibit a rapid decay of current within and beyond the active region, so that operation will not be affected by truncation of the structure. A major cause of the rapid current decay is, of course, the large radiation of energy from the active region. An additional cause, in at least some types of frequency-independent and log-periodic antennas, is the attenuation resulting from the rejection of incident energy by the reflection region (the filter stop-band effect mentioned previously). The prevalence and importance of this latter filter action are still uncertain.

Finally, two other observations may be made. Although we have tended to think of the structures of Figs. 4, 5, 7, and 8 as single antennas and the structure of Fig. 9 as an antenna array, it appears that most frequency-independent and log-periodic antennas may be thought of as antenna arrays, with the array factor playing an important role in the formation of a proper endfire or broadside pattern. The localization of the individual radiating elements may be easier to see for the case of the log-periodic dipole array of Fig. 9, but the array action can also be observed in the other cases; it is particularly evident in the case of the fairly tightly wrapped conical log-spiral.

The second observation relates to the similarity between antennas derived from the angle concept and log-periodic concept.¹⁷ Both lead to a solution of the unlimited-bandwidth problem and for this reason both have come to be known as frequency independent.

An example of the similarity between these two antenna types can be demonstrated in the case of the log-periodic wire antenna of Fig. 8, which produces a linearly polarized beam off the apex with the electric vector parallel to the transverse elements. If two such antennas are arranged in space quadrature along a common axis, and with a common origin but with one structure scaled a quarter period from the other, the resultant combination

produces a circularly polarized beam with a pattern that rotates about the axis with frequency, exactly as in the case of the conical equiangular spiral antenna. Conversely, of course, if the pattern of a conical equiangular spiral is probed with a linear receiving antenna of fixed plane of polarization, the measured pattern will vary log-periodically with frequency, as does the pattern of the antenna of Fig. 8.

In addition, it is pertinent to note that if a narrow-armed conical equiangular spiral (an angle structure) is flattened sideways (along the axis), it becomes a log-periodic zigzag antenna.

Recent developments

The log-periodic and angle concepts have been used to generate many highly useful antennas of large bandwidth. Fig. 17 shows a very practical two-element array of log-periodic dipole arrays capable of maintaining a nearly constant radiation pattern and a 50-ohm input impedance over the frequency range from 450 to 2000 Mc/s. The 50-ohm input impedance results from feeding two 100-ohm arrays in parallel. Although the dipole array is a balanced structure, it can be fed as shown with a coaxial cable running up the inside of one of the hollow transmission lines, utilizing the frequency-independent balun effect previously noted in connection with the conical log-spiral antenna.

Fig. 17. Two-element array of log-periodic dipole arrays.

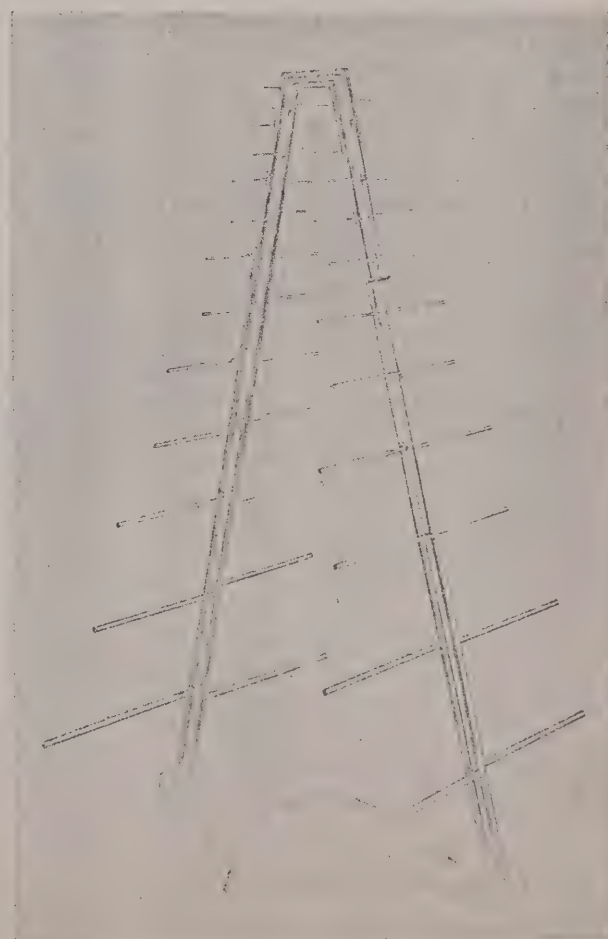


Fig. 18 shows one version of the log-periodic resonant-V developed by Mayes and Carrel.^{18,19} This antenna was designed to overcome one of the major shortcomings of the ordinary log-periodic dipole array—namely, the long physical length of array required to cover a very wide band of frequencies. The antenna of Fig. 18 is designed to operate in several modes. In the lowest order $\lambda/2$ mode, the operation is similar to that of the log-periodic dipole array because the forward tilt of the elements has small effect for this mode. However, as the frequency of operation is increased beyond that at which the shortest elements are resonant—that is, when the active region runs off the front end of the array—the largest elements at the rear become active in the $3\lambda/2$ resonance mode. In this mode the forward tilting of the elements ensures a good unidirectional pattern of high directivity. As the frequency is further increased, the active region moves forward through the array in the $3\lambda/2$ mode until once again it runs off the front end, to return to the rear in the $5\lambda/2$ mode. This scheme makes it possible to obtain large bandwidths of the order of 20 to 1 with a relatively compact array. The pattern and impedance characteristics remain good over the entire frequency spectrum except for intervals about the mode-transition frequencies. Based on these principles, arrays have been designed to cover all of the television channels from 2 through 83, corresponding to a frequency range from 54 to 890 Mc/s.

Another interesting development is that of a log-periodic folded-dipole array. At first thought it would appear that such an array could not work because the short elements at the front of the array present a very low impedance, thus short-circuiting the transmission

region leading to the active region. This difficulty is circumvented²⁰ by connecting the folded dipoles in series with the transmission line, rather than in shunt, and recognizing that the active region will occur near first resonance, that is near the element length ($\lambda/4 < l < \lambda/2$) where the capacitive reactance of the short antenna resonates with the inductive reactance of the folded dipole viewed as a short-circuited transmission line. This unusual operating mode for the folded dipole results in a shorter element length for resonance, and consequently a narrow width for the resulting folded-dipole array.

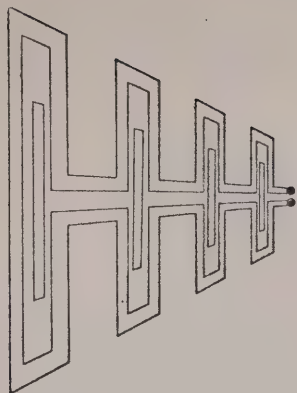
A major problem with log-periodic structures has been the design of an antenna that will operate successfully when fed against a ground plane to produce vertical polarization. One half of the antenna of Fig. 8 can be operated over ground to produce horizontal polarization, as can an inclined horizontal log-periodic dipole array. For vertical polarization, particularly in the high-frequency band (3–30 Mc/s), it is desirable to use the equivalent of a log-periodic monopole array that has a height of only approximately $\lambda/4$ at the lowest operating frequency, rather than $\lambda/2$. Because of the necessity for introducing a transposition between elements (or otherwise producing the required phase difference between elements) it is not possible simply to use one half of a log-periodic dipole array fed against ground.

Several solutions to this problem, having varying degrees of success for different applications, have been developed by a number of workers in the field.^{21–23} A quite recent development²⁴ using folded monopoles with added phasing elements promises to be very useful.

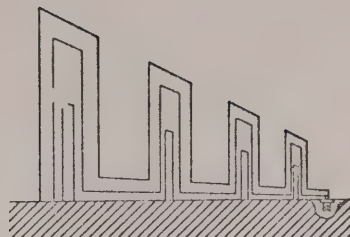
Three versions of this antenna are shown in Fig. 19.

Fig. 18. Log-periodic resonant-V array, for operation in several modes.

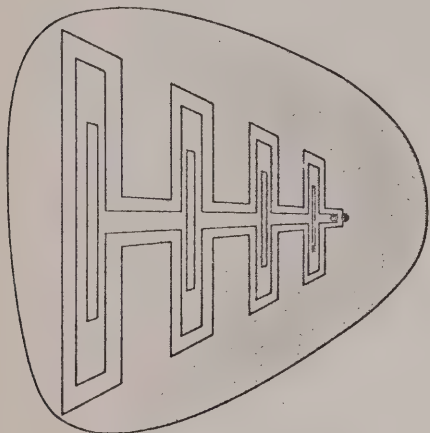




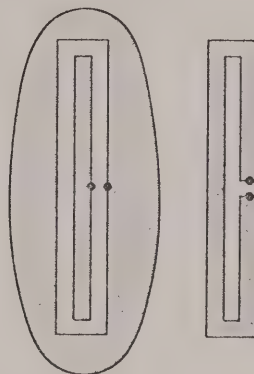
A



B



C



D

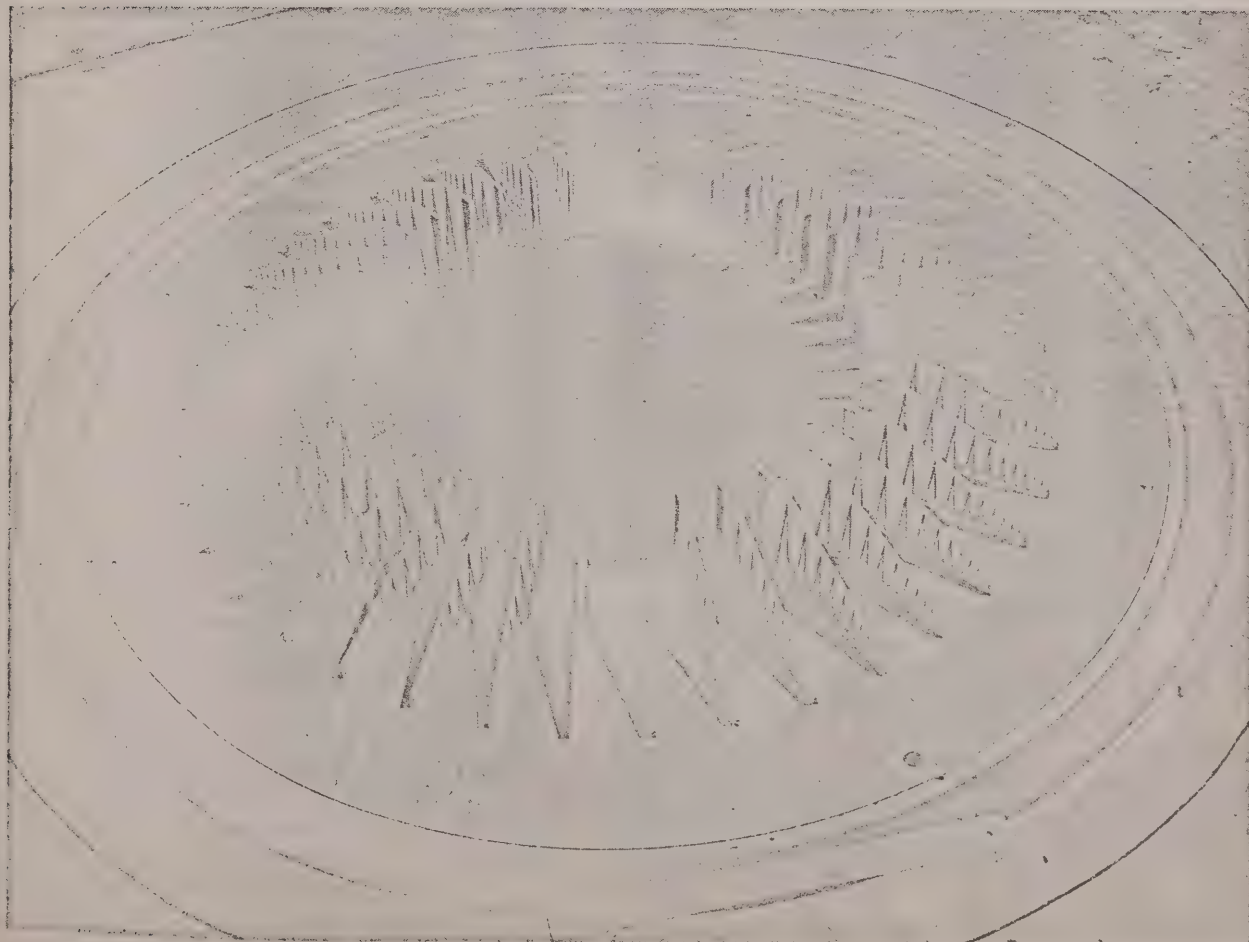


Fig. 19. Log-periodic arrays of folded elements.

A—Log-periodic folded-dipole array. B—Log-periodic folded-monopole array. C—Log-periodic folded-slot array. D—Duals: folded slot and folded dipole.

Fig. 20. Model of wide-aperture log-periodic array for high-frequency radio direction finding (3–30 Mc/s).

The log-periodic folded-slot array (A) was conceived first, but by duality, the log-periodic folded-dipole array (B) is obtained automatically. Because this array possesses the proper image symmetry about the horizontal axis (horizontal currents in opposite directions, vertical currents in the same direction), one half of the array can be fed against a ground plane to produce the folded monopole array of (C). The duals, folded slot and folded dipole, are illustrated in (D). The dimensions of the phasing slots in (A), or phasing strips in (B) and (C), are adjusted experimentally to provide the required phasing between successive dipoles or monopoles to produce a good backfire beam.

For greater directivity than can be achieved with a single frequency-independent antenna (or array) it is possible to use the frequency-independent structure as the broadband feed of a large paraboloid. Although the resultant combination is no longer frequency independent, high-gain antennas having a usable bandwidth as high as ten to one have been built by use of this approach.

Some of the high-gain paraboloid tracking antennas for the Atlantic Missile Range have been modified to use two conical log-spiral antennas as a circularly polarized broadband feed in a conical scan system. This application covers a frequency band of 215 to 1000 Mc/s, but the feed elements themselves are capable of operating continuously to 2300 Mc/s.

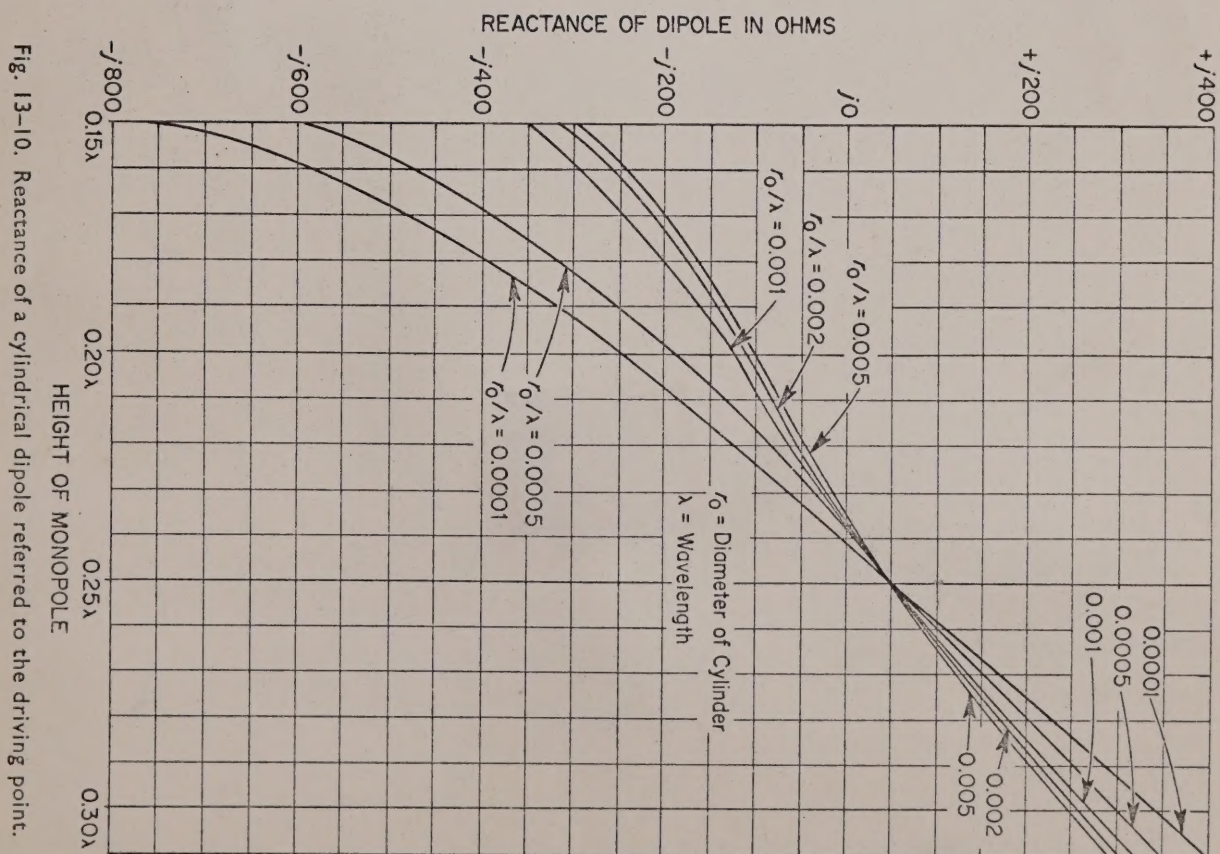
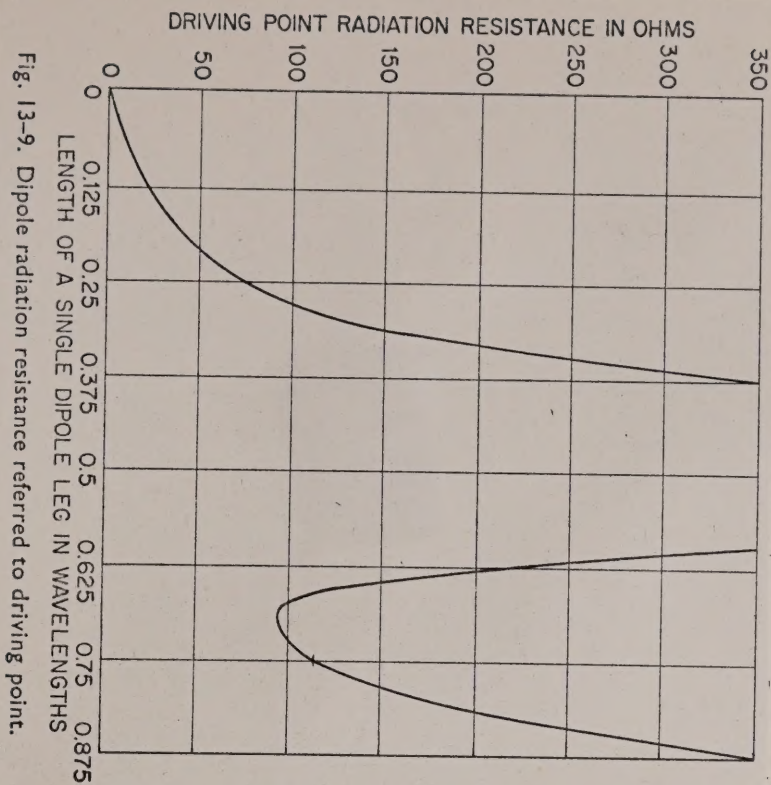
An alternative approach to the high-gain broadband problem is illustrated in the model of a broadband (3–30 Mc/s) wide-aperture radio-direction-finding array shown in Fig. 20. For frequency-independent arraying, the individual elements should lie along radials and be arranged to fire inward toward the common origin (toward the hub of the wheel). Unfortunately this arrangement requires opposite elements to fire through each other, and severe pattern deterioration results. In the array of Fig. 20 the log-periodic antennas fire outward. A 100° sector of elements is connected together through an appropriate phasing network and rotating switch or goniometer to form a narrow beam, which rotates with the goniometer as the latter connects in elements on one side of the sector and disconnects them on the other side. Again, this arrangement is far from being frequency independent, but the use of broad-band log-periodic structures as array elements is an improvement over the earlier use of frequency-sensitive elements.

This last example indicates that although truly remarkable progress has been made in the past decade in achieving broadband antenna operation there still remain some challenging problems for the future. Among these challenges are the design of broadband antennas having very high gain, and the design of frequency-independent antennas to produce specified radiation patterns.

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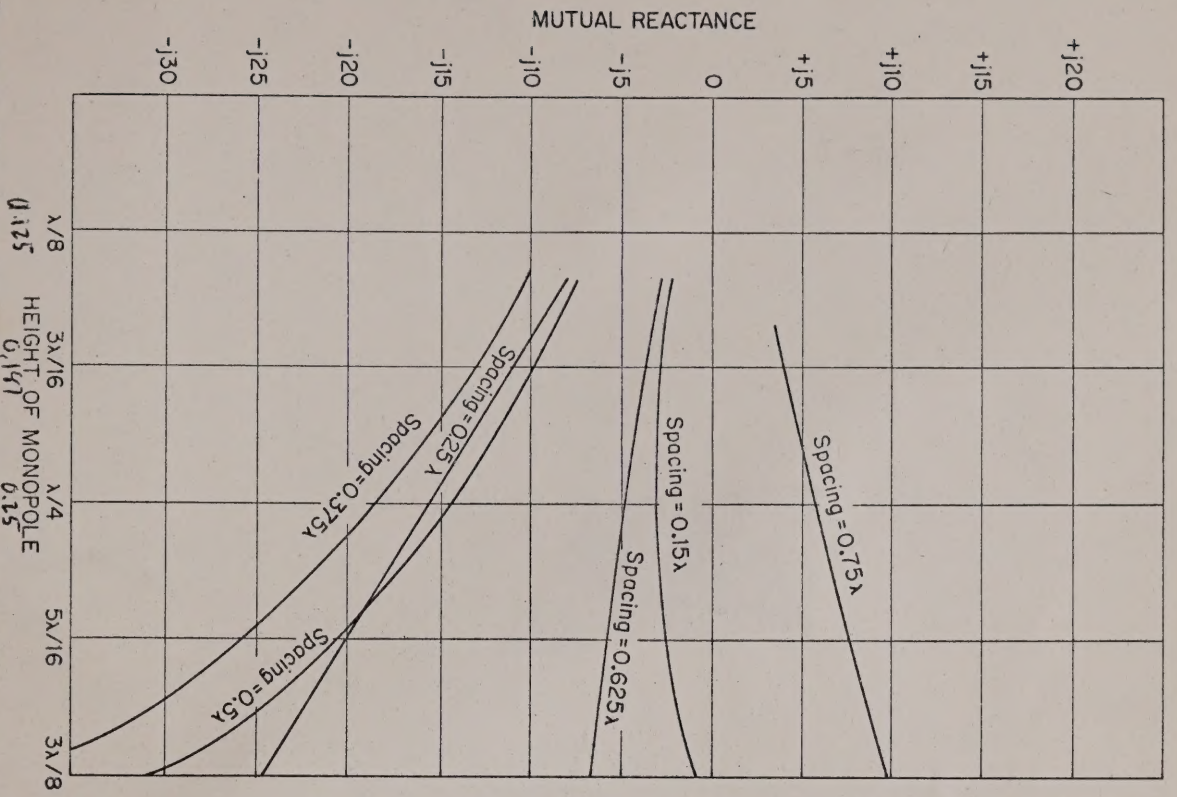


Fig. 14-12. Mutual reactances between a quarter-wavelength monopole and other length monopoles for various spacings.

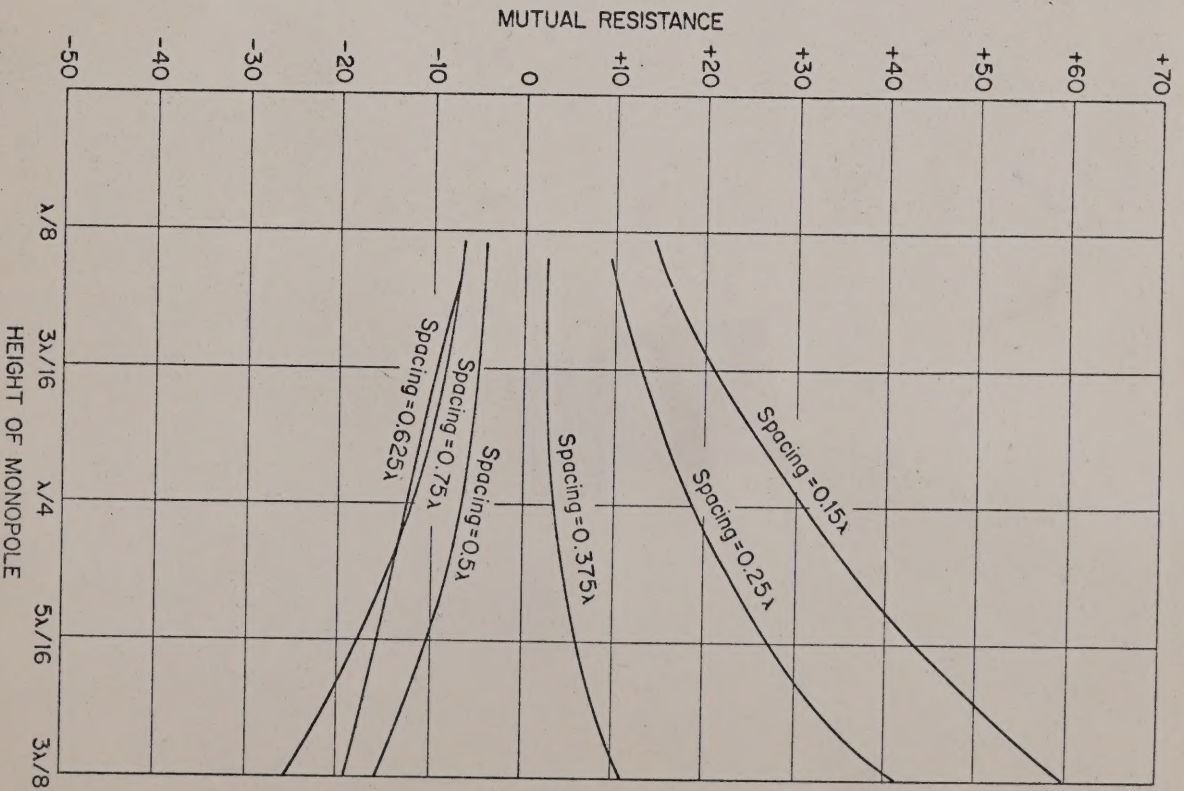


Fig. 14-11. Mutual resistances between a quarter-wavelength monopole and other length monopoles for various spacings.

